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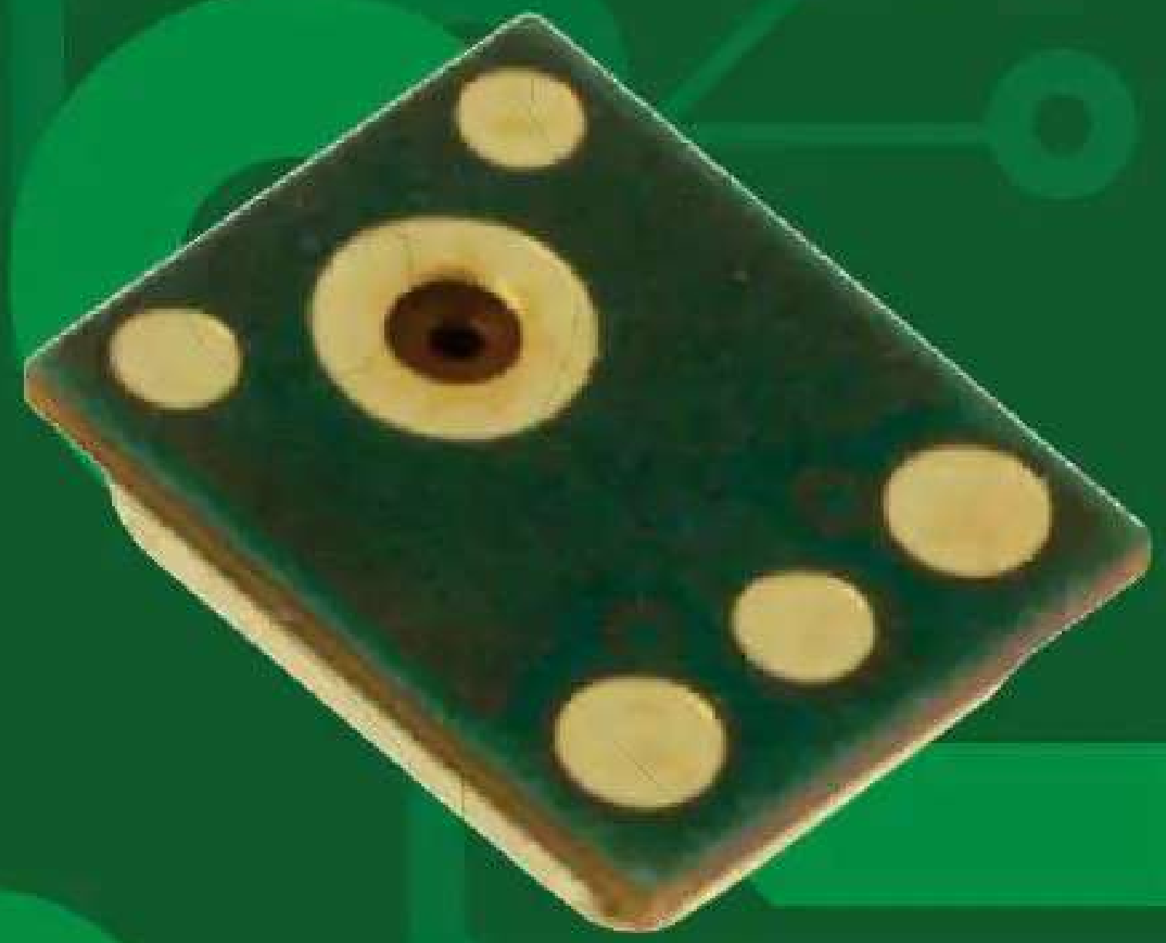
Audio Out
Getting the best out of capacitors for HiFi, pt2

Precision Electronics
Analog signal switching techniques and tips

ESR 3.2V - - -
0.02Ω
S1
S2
S3

ESR Test Tweezers
Measure capacitor equivalent series resistance

A quiet domestic Fan Speed Controller



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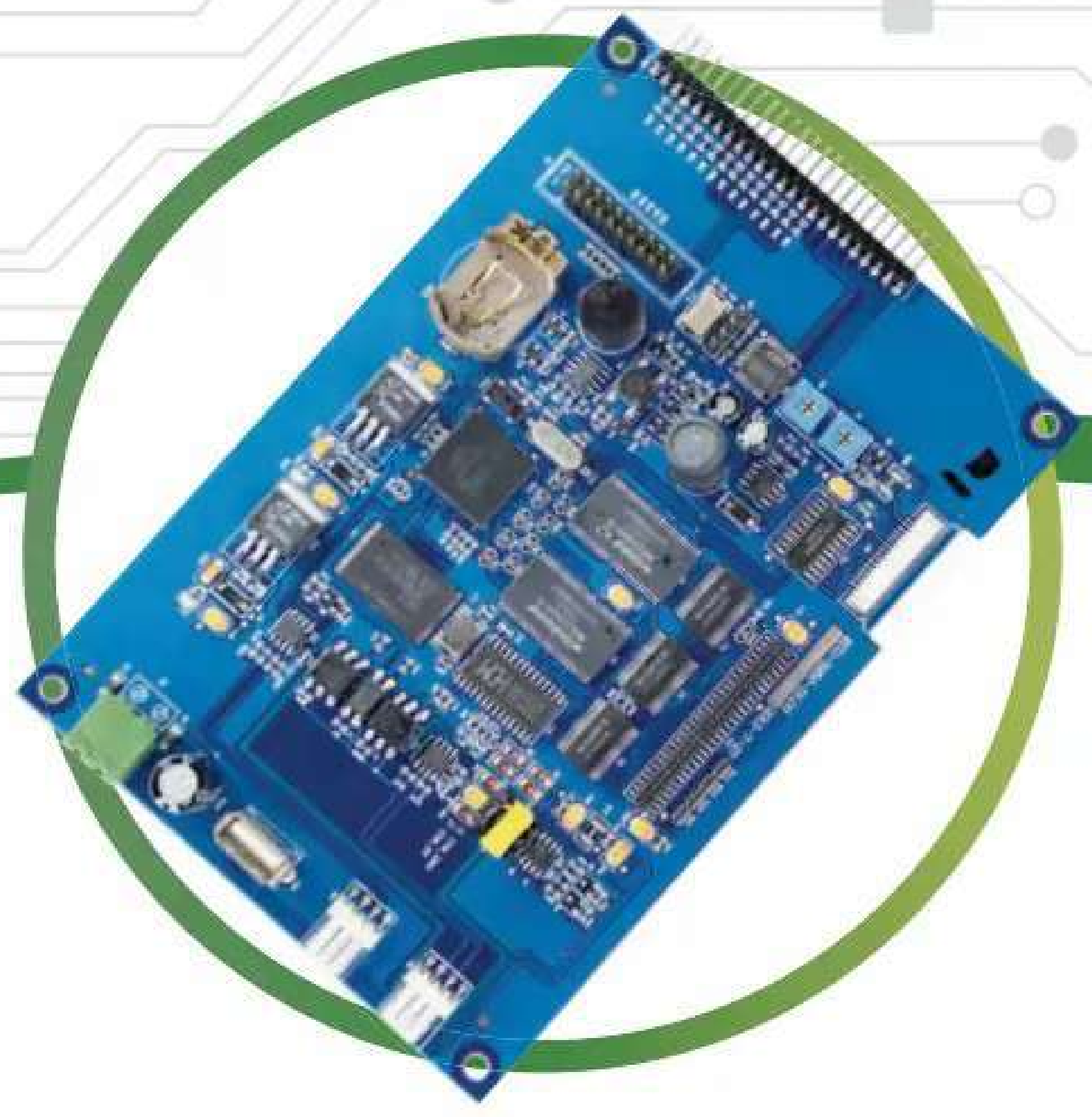

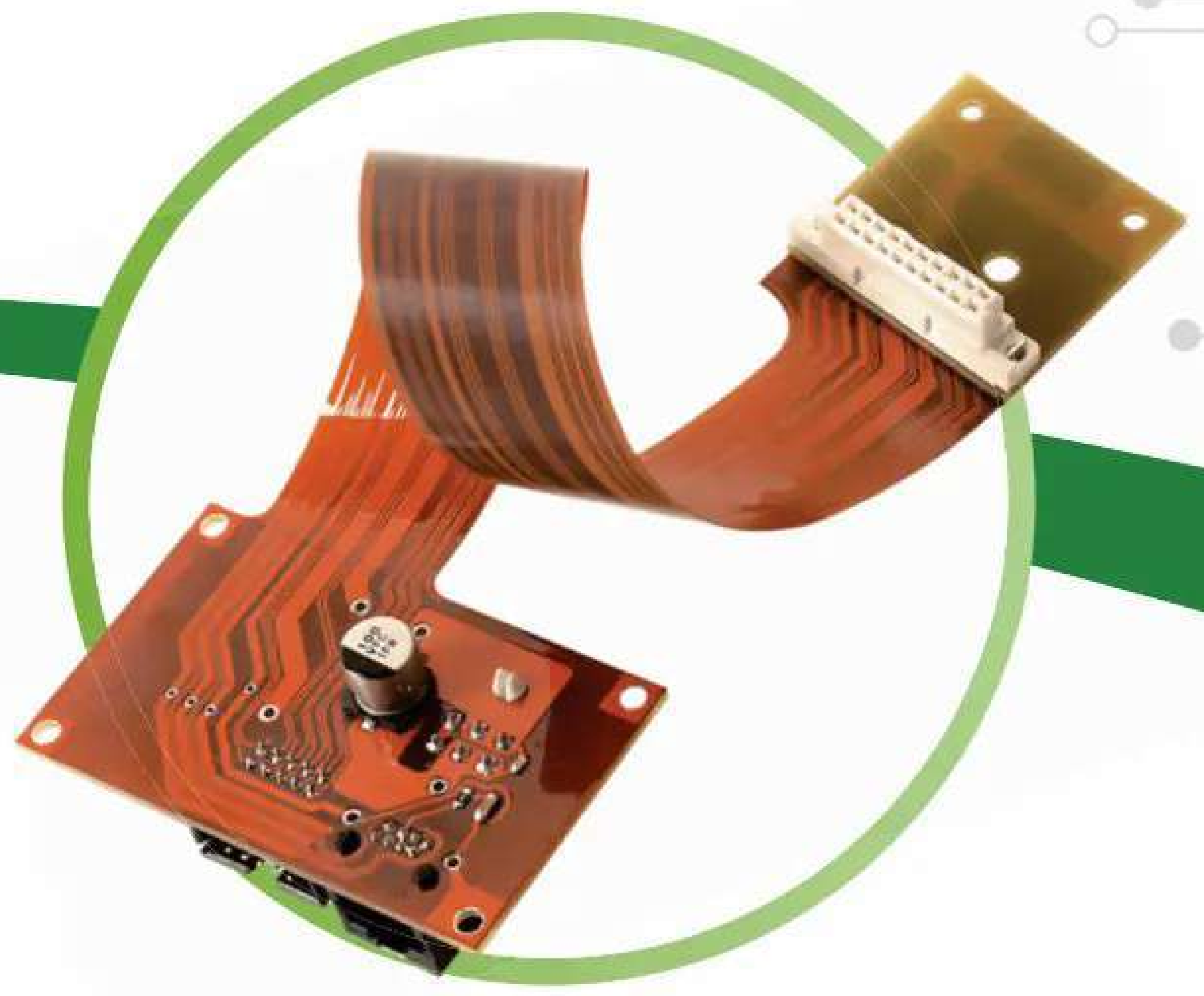

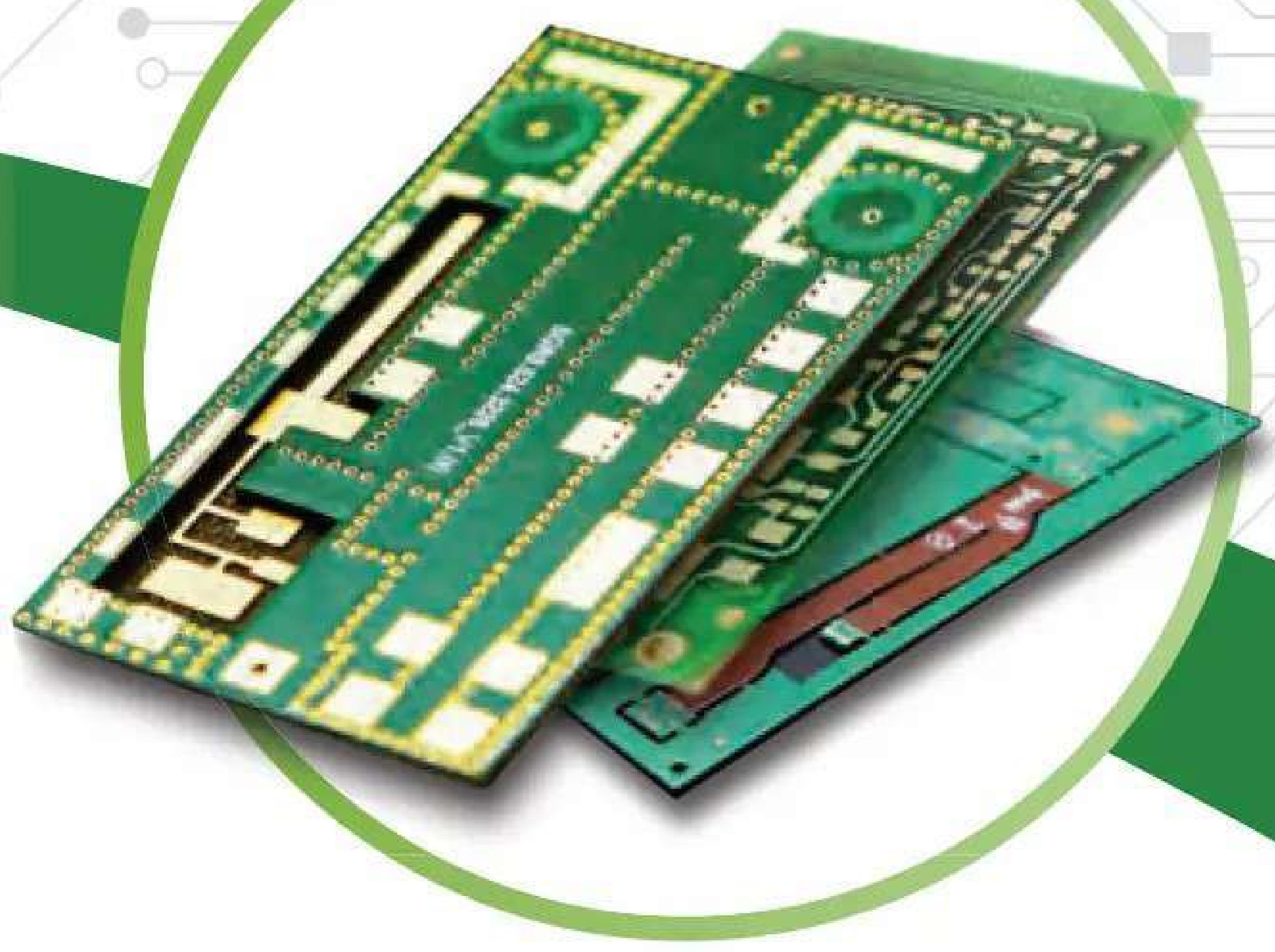

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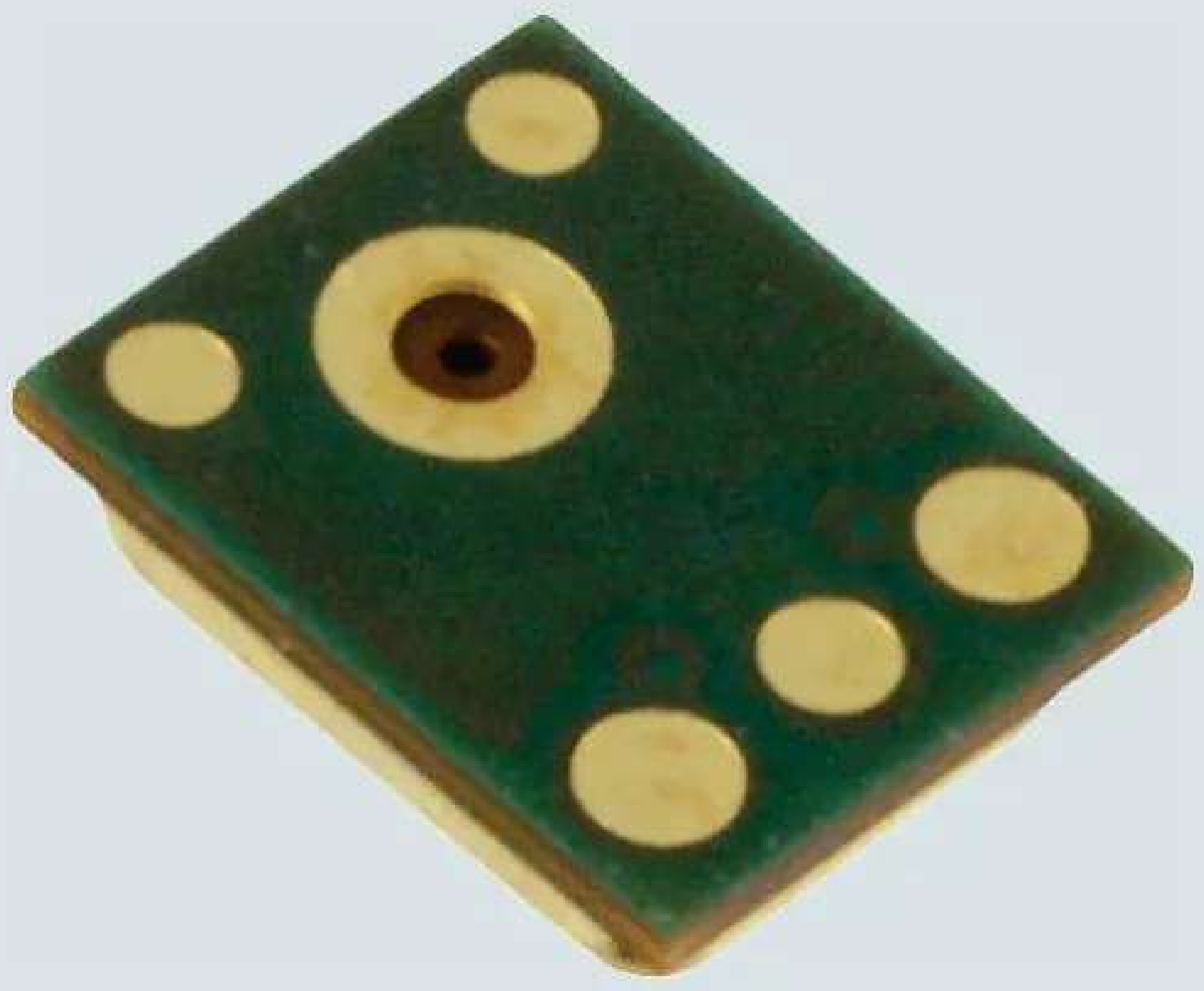

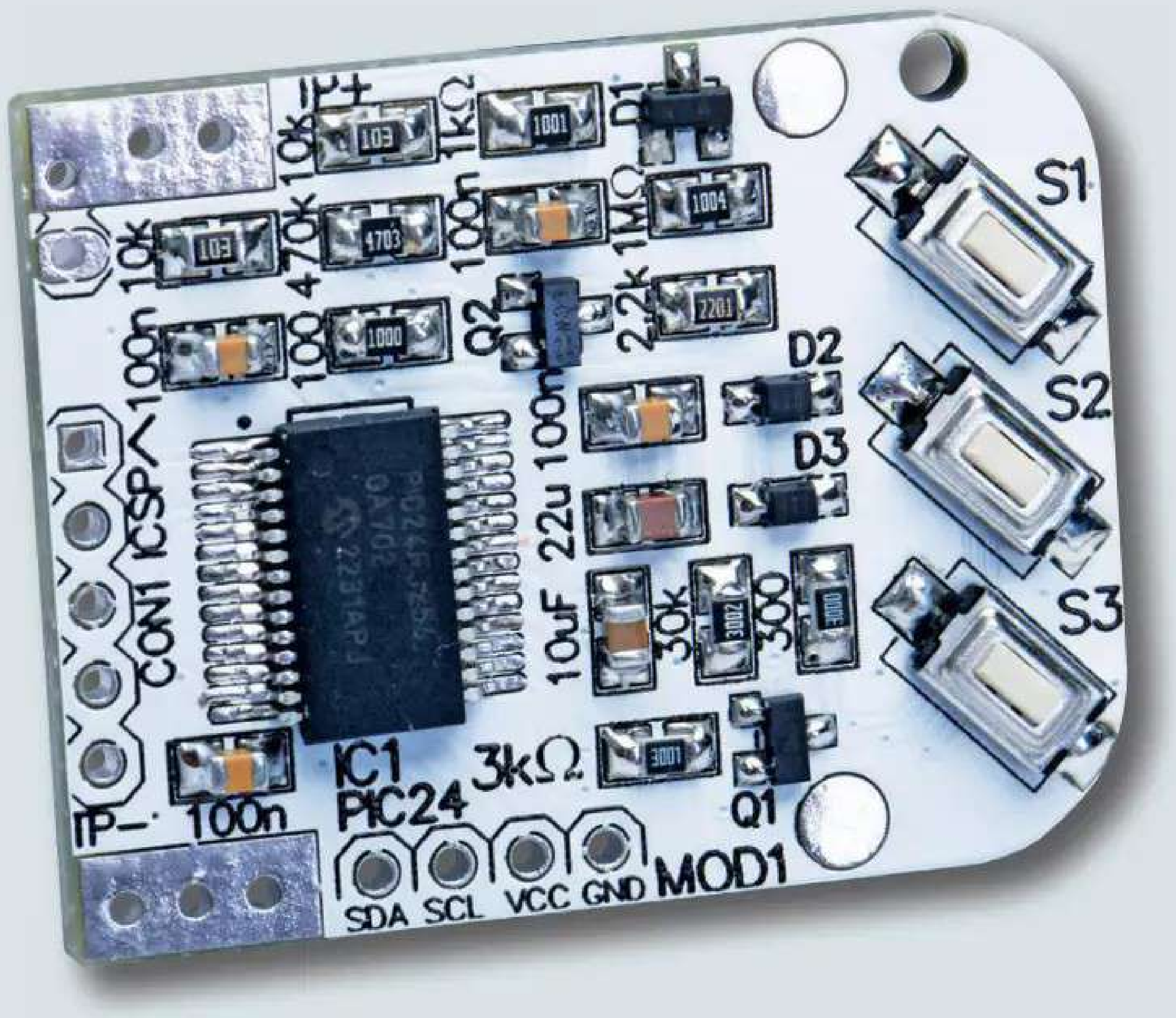

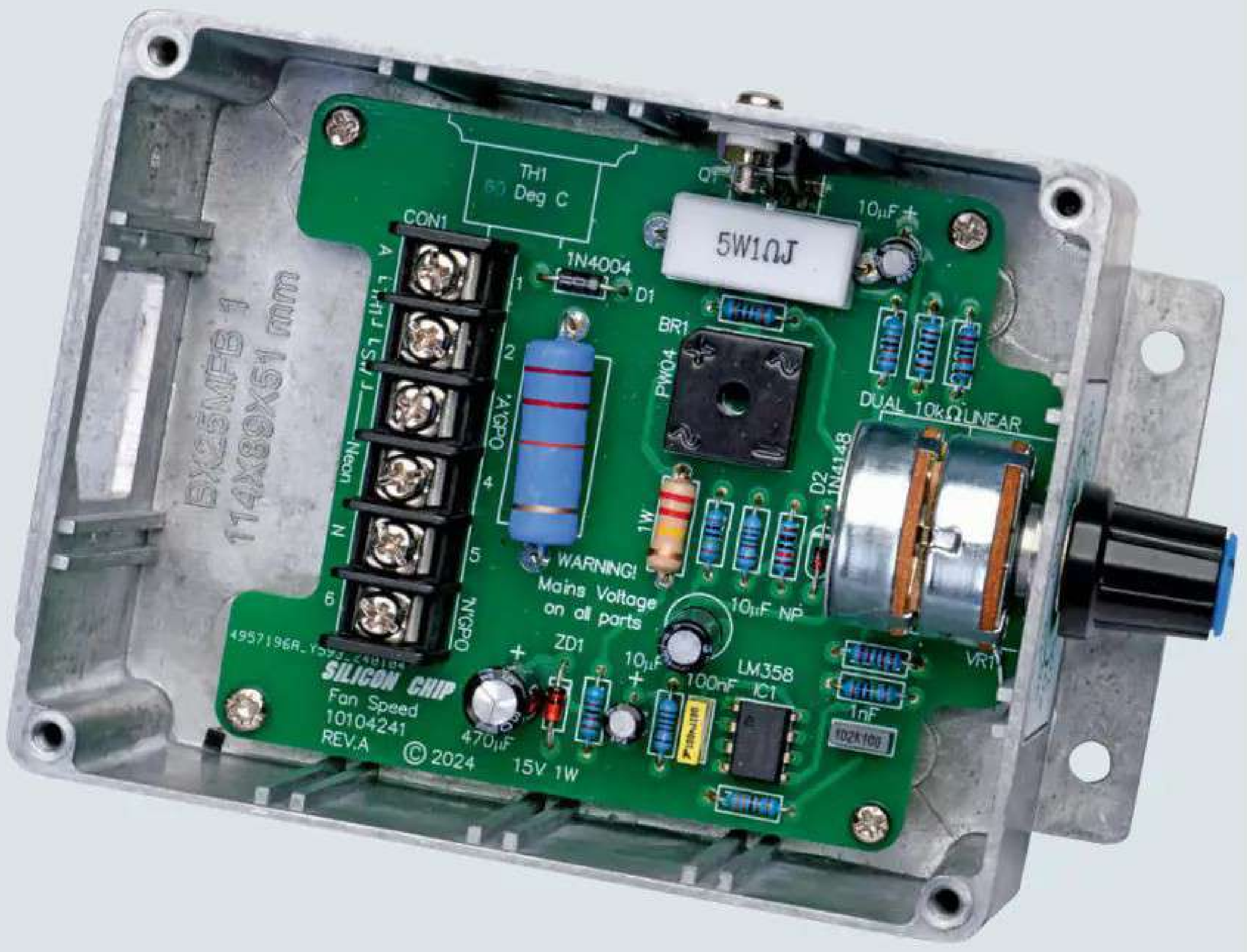

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Equivalent Series Resistance testers are very useful

One of our projects this month is the very useful ESR Test Tweezers tool, a handheld device that can measure capacitance (over a limited range) but, more importantly, can also accurately measure capacitor equivalent series resistance (ESR) down to tiny fractions of an ohm.

I was a bit surprised when I looked back over the recent history of *Practical Electronics* and couldn't find much in the way of ESR testers until our recent LC/ESR Meter from the August 2024 issue. Given how useful these devices are for repairing and servicing equipment, I expected other designs to have been published.

For example, SILICON CHIP magazine published Bob Parker's ESR Meter Mk2 design in its March & April 2004 issues. It turned out to be an extremely popular project that was made into a kit by both Dick Smith Electronics and Altronics.

The beauty of an ESR meter is that it can identify faulty electrolytic capacitors that otherwise seem fine. Electrolytic capacitors are probably the single most common source of faults in equipment these days as they have a much shorter lifespan than any other component, usually measured in thousands of hours.

An electrolytic capacitor might look fine, and a capacitance meter may even give the expected reading. However, if it has developed a high ESR (often as a result of the electrolyte drying up or leaking out), it won't properly perform its job of filtering or smoothing out voltages. An ESR meter will immediately detect this and pinpoint the capacitor(s) that need replacing.

It is not uncommon to find a large electrolytic capacitor that seems fine on visual inspection and would normally have an ESR figure of 0.01Ω or less, that then measures several ohms! Such a capacitor is essentially no good for anything other than decoration at that point.

Of course, if a capacitor has a bulging top or a brown stain around its base, you probably don't need any test equipment to tell you it needs replacing. Unfortunately, faults aren't always that obvious. That's why an ESR meter is a worthy addition to a repairer's toolkit.

Errata

Unfortunately, some errors in the Audio Out column last month were not spotted before it went to press:
- In Fig.15 on page 24, the 1.5nF capacitor should be 1.5µF and the 47nF capacitor should be 4.7µF.
- In Fig.22 on page 25, the 2.2nF capacitor should be 2.2µF.

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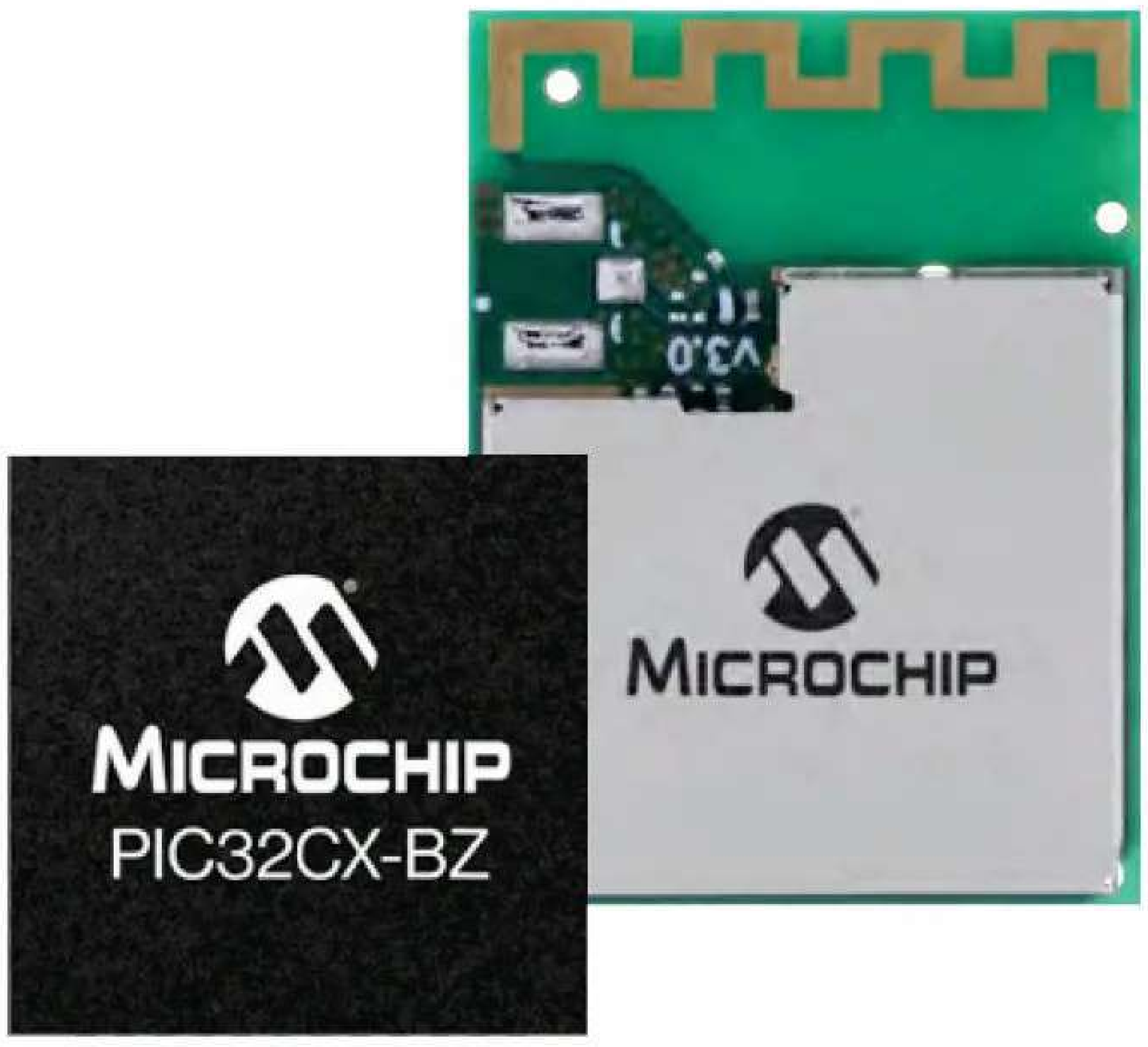

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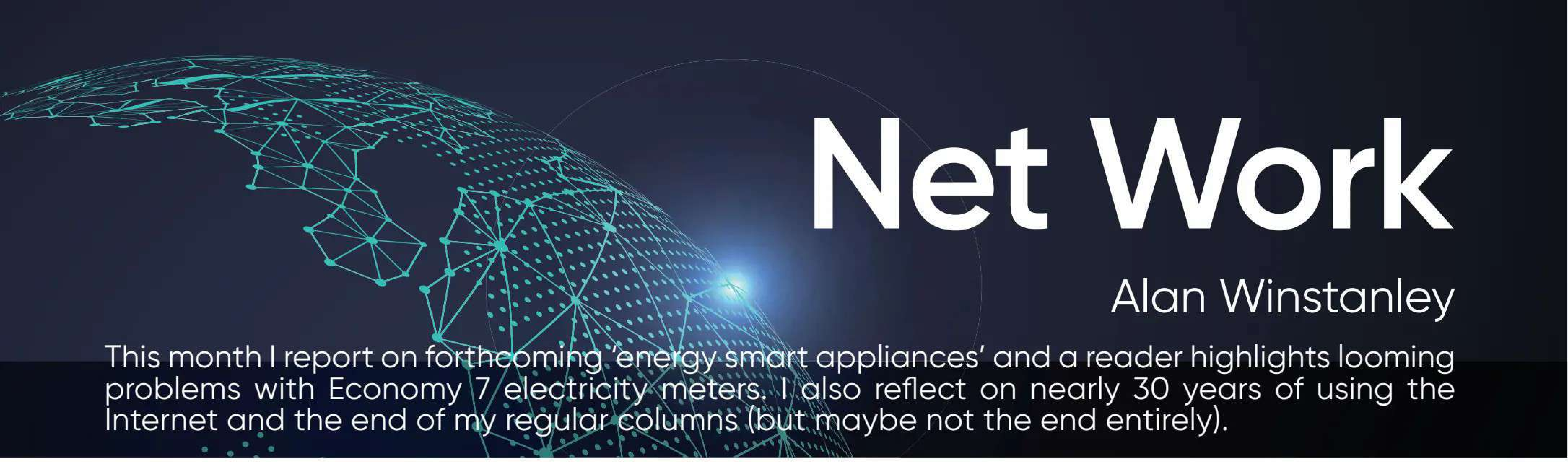


Net Work

Alan Winstanley

This month I report on forthcoming 'energy smart appliances' and a reader highlights looming problems with Economy 7 electricity meters. I also reflect on nearly 30 years of using the Internet and the end of my regular columns (but maybe not the end entirely).

Smart meters have been a hot Net Work topic for several years; the energy sector continues its drive to install them in every British home. There's no doubt that many consumers appreciate having a smart meter's in-home display (IHD) to track energy consumption, but BBC News reported that about four million of them currently don't work properly.

Much has been written about smart meters going 'dumb', having connectivity problems or sending inaccurate user data, resulting in overcharging.

Some householders are even being paid to accept one: EDF (Électricité de France) recently offered the author a £150 bounty to sign up for a smart meter, claiming that my current digital electricity meter has reached the end of its life. This novel twist may finally tempt customers into accepting the latest SMETS2 (Smart Metering Equipment Technical Specifications) meter.

I wrote half a decade ago (*Net Work*, January 2020) about how the concept of a Smart Meter Home Area Network (SMHAN) would forever alter how householders used appliances such as dishwashers, tumble dryers or electric vehicle chargers.

Industry and the government are now talking about a new generation of 'energy smart appliances' (ESAs) as a way of managing Demand Side Response (DSR) – defined as "a means of changing electricity consumption **in a way that benefits the electricity system**".

ESAs have yet to materialise, but a voluntary EU Code of Conduct for 'interoperability and energy smart appliances' was adopted Germany last year, and the technical standards are now being hammered out.

Once connected to a household's SMHAN, energy smart domestic appliances would 'know' what the cheapest tariffs are at any time of day, and optimise the consumer's energy usage and billing accordingly. This may well appeal to EV owners especially.

Appliances are getting 'smarter'

To see which way we're heading, the British Standards Institute offers heavy reading in a publicly available specification (PAS) in PDF format at https://pemag.au/link/ac4h

I was gratified to read that the proposals specify a "…standardised control, **subject to an explicit consumer consent**, of energy smart appliances (ESAs) on an electricity network." So consumers would have to opt in to take advantage. Like smart meters, ESAs will probably be touted as offering a consumer benefit, but the truth is that they're all about rationing our faltering electricity supply.

According to Electrolux, nine manufacturers in addition to Electrolux Group – including Miele, Turkey's Arçelik, Mitsubishi Electric, Panasonic and Vaillant Group – have committed to the EU Code of Conduct and they promise to develop 'interoperable connected products' within a year, including washers, tumble dryers, dishwashers and heating/ventilation/air-conditioning (HVAC) systems.

My guess is that by the end of the decade, consumers will routinely sleepwalk into accepting this method of silently controlling their electricity consumption. Still, we will have to see whether ESAs will make much difference to our energy supplies.

Not so economical

There are problems looming for legacy meter users as well. Regular reader **David Platt** dropped me a line about another energy-saving problem he's wrestling with:

Being a retired coal power shift operations engineer, I follow your comments about the future progress of new energy generation with great interest, especially the advancement of SMRs, wind, solar and wave power.

I have what is known as a "GEC Radio Teleswitch" (RTS) feeding a Siemens Normal/Low metering device. Are you aware that in June 2025, Octopus Energy is cutting off the Economy 7 radio signal, which is what my meter requires to operate properly? They propose giving me a new-fangled dumb smart meter instead.

Is this signal cut-off common for all Energy Providers, or is this just a way to roll out the Government's dumb smart meters? Can we stop the signal cut-off? Yours, David Platt MIET.

The concern David highlights relates to the abandonment of the 1980s RTS long wave transmission, which takes effect on the 30th of June, after being postponed from 2024. The LW radio signal technology is deemed obsolete and unserviceable, so naturally all the electricity suppliers (not just Octopus) are pushing smart meters as its replacement.

Some 600,000 meters are reportedly affected, and there isn't really any technological alternative on offer today.

For non-UK readers, Economy 7 and Economy 10 are cost-saving electricity tariffs that use a radio-controlled device to switch certain meters to cheaper off-peak rates, for seven or ten hours. There is more information about the RTS changeover at https://pemag.au/link/ac4i

As far as I know, having a smart meter fitted still isn't compulsory, but if you use Economy 7 or 10, you will otherwise lose the cheap-rate tariff once the RTS is switched off. Unless they have a change of heart (which I doubt), I see no hope of extending the service. I'm afraid the outlook isn't great for consumers who don't want a smart meter.

If no smart meter is fitted, the end user must bear the cost of ripping out the old equipment and updating his wiring. David later wrote to say that he had resigned himself to scouting around for a smart 'dumb' meter after all.

When ill winds blow…

We now have the ludicrous situation of sometimes paying wind farm operators not to generate power (shedding load) when it's too windy, while paying end-users of the so-called demand flexibility service (DFS) not to use electricity at peak times either.

There are signs that the novelty has worn off this incentive scheme as, according to the *Telegraph*, only 750,000 consumers remain in it compared with two million in the previous year.

Britain's costly quest for renewable energy is fraught with setbacks. Last December, the brand new 190-acre, 50MW Porth Wen Solar Farm in Wales, which was six years in the planning and construction, sustained severe damage caused by Storm Darragh and is already out of service. A nearby wind turbine also self-destructed because it was too windy.

Such turmoil is likely to get worse over the next few years, and I reckon it's maybe a decade away before Small Modular Reactors finally come on stream to provide a stable base load. EDF recently confirmed that two of its four ageing Advanced Gas-Cooled Reactors will now continue operating until 2027, and the remaining two can operate until 2030.

The extra toll that electric vehicles and power-hungry AI data centres will place on the grid has yet to be played out, maybe helped by energy-smart appliance technologies. No-one knows for sure if wind and solar renewables will take up the slack in the meantime.

For my part, I've stocked up on refills for my picnic stoves and portable heater, and I'll keep a cupboard stocked with tinned food and bottled water just in case!

Going full circle

The first *Net Work* column in August 1996 was just a single page, written to support users grappling with a newly emerging fad called the Internet. The buzzwords 'Cyberspace' and 'Information Superhighway' had entered our lexicon (they've pretty much left since), but home PC and laptop ownership had yet to hit the mainstream due to the high cost and complexity.

The early Internet was mainly the province of computer enthusiasts, hobbyists, academics and professional types.

For many 1990s home computer owners in the UK, American company CompuServe Information Services offered their first taste of going online. It was a 'walled garden' of CIS forums and correspondence via email using the CompuServe messaging system. A 1990s video titled The Kid's Guide to the Internet was a nostalgic (if cringeworthy) intro that captured the zeitgeist 30 years ago. It's on YouTube at **https://youtu.be/mfMrVKnGzwg**

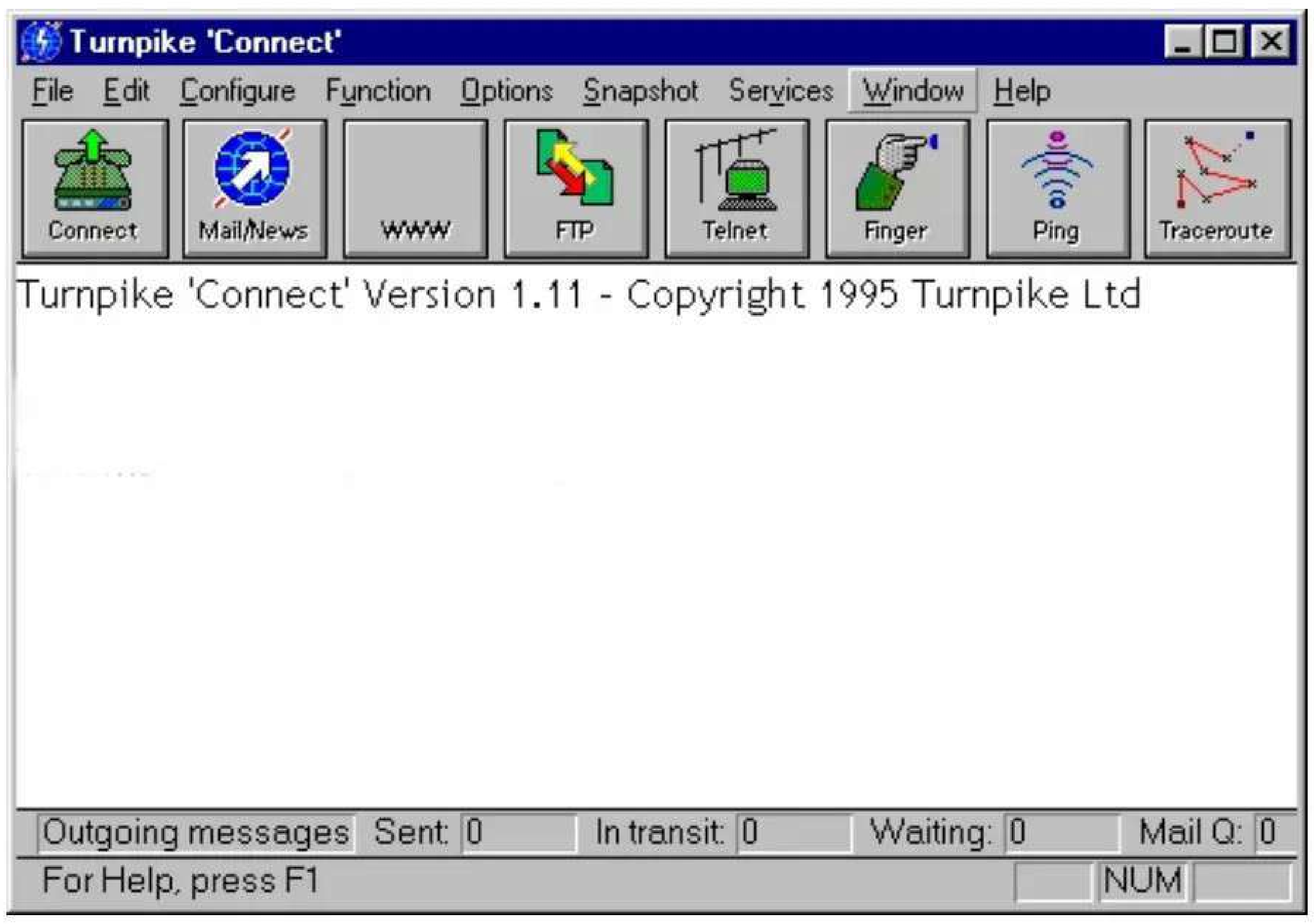
After a rewarding period of authoring using *Locoscript* on an Amstrad PCW9512 word processor, the IBM-compatible PC beckoned me and, in 1993, a new Ambra 486 PC arrived, with an internal 14.4kbaud modem providing dial-up access.

I could also upload my monthly magazine 'copy' to the publisher by direct modem-to-modem transfer, using *Procomm Plus* for Windows 3.1, which saved me from having to post off a floppy disk.

In Britain, a fixed-rate 'raw' Internet service was first supplied in the 1990s by London-based Demon Internet, in the shape of their 'tenner a month' (TAM) account. For £10, 'Demonites' could enjoy unlimited online access, and so the author's **epemag** moniker was first created. The phone bill was literally ¼-inch thick, listing 0845 telephone numbers to Demon's nearest point-of-presence (PoP).

Early Windows TCP/IP communications software barely existed, and no-frills, text-only DOS implementations such as KA9Q were provided by Demon Internet (see **https://en.wikipedia.org/wiki/KA9Q**).

A new 50MW solar farm at Porth Wen, Wales suffered major storm damage in December and is out of action. A wind turbine was also destroyed. (Source: YouTube/NWSXS).



Turnpike 'Connect' Version 1.11 - Copyright 1995 Turnpike Ltd

Outgoing messages Sent: 0 In transit: 0 Waiting: 0 Mail Q: 0

For Help, press F1

Demon Internet's Windows TCP/IP software included a dialler, network tools, an email client and Usenet newsreader.

It was slow going; a 5MiB file took half an hour to download, assuming the line didn't drop and one had to start all over again. To save time and money, 'zipping' large files was obligatory, and arcane techniques such as UUEncode made transferring large files more feasible.

Here is the news

One key online attraction was Usenet or newsgroups, a system of static forums dedicated to a particular topic such as *sci.electronics* or *comp.robotics.misc*. It's where I bumped into Clive (Max) Maxfield, who at the time was heading out from England to work in the USA, and was a big fan of *Practical Electronics*. The outcome of our chance encounter is evident in this month's issue!

Mainstream Usenet morphed into Google Groups, support for which was discontinued in 2024 (**https://groups.google.com**). Usenet still has its fans, and several paid-for services are sold online. Usenet was also an open source of email addresses, which fuelled the first spam campaigns, and the scourge of junk mail started to fill our mailboxes.

If nothing else, these tentative steps were a promising start, and then in 1995, new Turnpike for Windows software was launched. This program centralised a dialler, email, Usenet and other tools, making Internet access much more productive. Turnpike is described on Wikipedia at **https://en.wikipedia.org/wiki/Turnpike_(software)**

With user numbers increasing exponentially, dial-up speeds edged up to 28.8k baud, then 33.6k and finally 56k, the maximum rate realistically achievable over copper wires. The demand for Internet access led to more British households ordering second lines, but there simply weren't enough phone lines to go round.

The temporary solution was DACSing (Digital Access Carrier System) as a form of 'line splitting', which British

Inside Amazon's first book warehouse in Seattle. Founder Jeff Bezos being interviewed for TV by Robert Cringely in 1998.

Telecom introduced to provide 'virtual' phone lines, as noted in the Wikipedia page at https://en.wikipedia.org/wiki/Digital_access_carrier_system

Home Internet users feared their dedicated modem line being DACSed because dial-up speeds could drop by half, but BT were only legally obliged to provide a dial tone for voice calls, with no guarantee of data throughput. Mercifully, I escaped the dreaded DACS.

A web of intrigue

Early on, the world-wide web (WWW) consisted of a smattering of websites and, compared with email, Usenet or bulletin boards, the web was a lesser application for a while. Internet magazines sprang up that carried free cover-mounted booklets listing the latest 'must-see' URLs. There were even paperback book directories of website addresses!

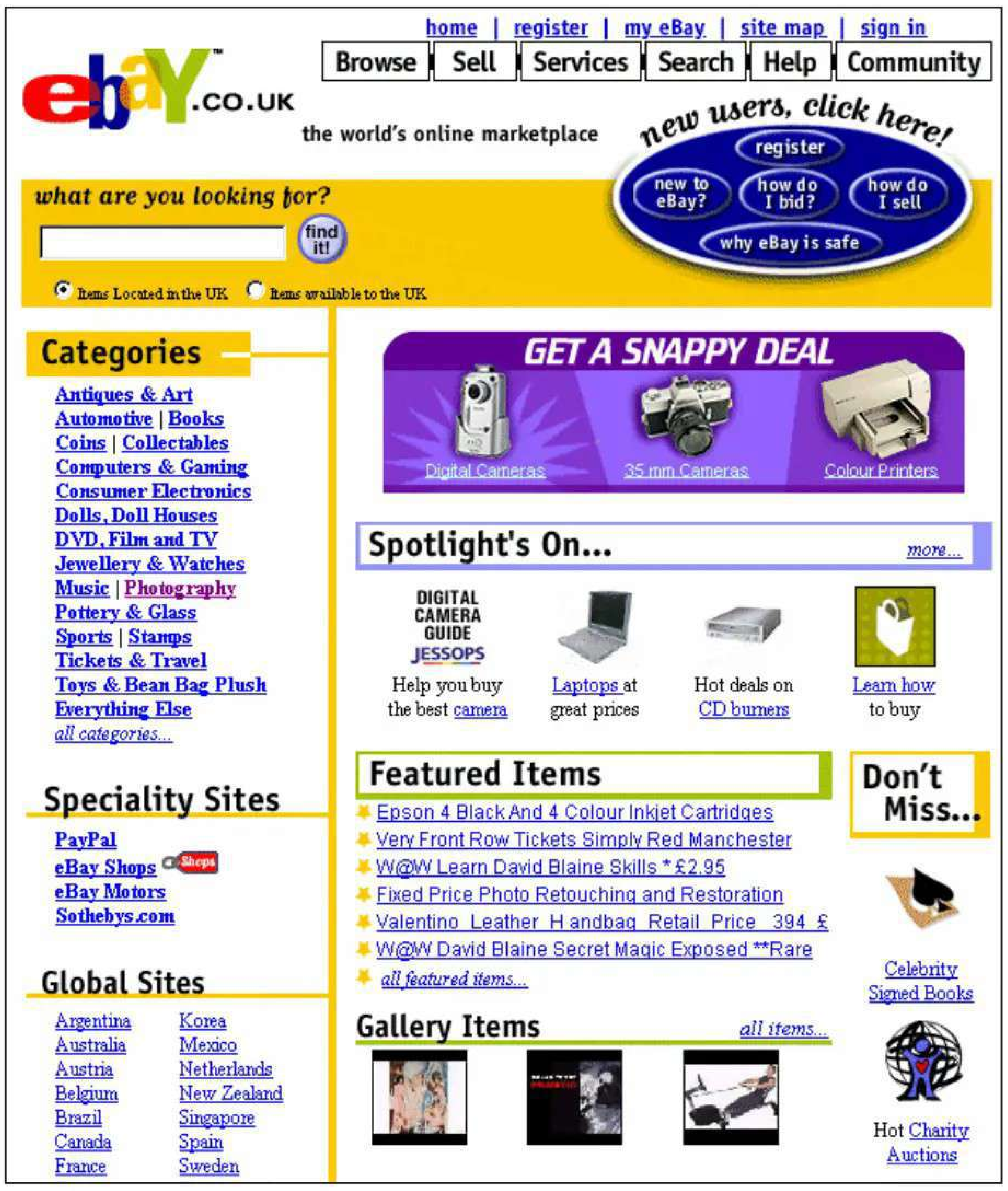
The idea of buying and selling on the web also took hold, although the UK was held back in this area by the dearth of available online payment processing services. Then an online bookshop called Amazon sprang up.

Speaking in a 1998 TV interview with Robert Cringely, Jeff Bezos recounted how, in 1994, he learned that Internet usage was rocketing by 2300% per year, so he short-listed 20 products that he thought he could sell online.

He chose to sell books; his logic was that so many book titles were available that it was impossible to list them in, say, a traditional printed catalog, so a searchable website would be the perfect solution. He then rented a Seattle warehouse and staffed it with Gen. Z book lovers, and Amazon started shipping books all around the world.

In the mid 1990s, I had a whale of a time importing books at prices much lower than those seen in the UK, sometimes delivered before titles were even released in Britain. Initially, Amazon burned through cash and made heavy losses. Deciding whether Amazon would even stay in business became a sport amongst the commentariat.

Amazon defied all the odds and expanded its range ("Everything from A to Z"), setting up shop in the UK and Germany in 1998, and the rest is history.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Early eBay's website was tightly designed and easy to navigate – unlike the bloated one of today.

A Oh No?

Another turning point occurred in 1994, when America Online (AOL) users started to flood onto Usenet, trampling on established courtesies and usage protocols along the way ("Me, too!" was their often-mocked catchphrase).

AOL floppy disks and CDROMs rained down in a way we'd never seen before, in an attempt to batter potential customers into signing up to AOL's fixed-rate service.

As described in the fascinating Anglo-American book *net.wars* by Wendy M. Grossman (New York University Press, ISBN 0-8147-3103-1), AOL also provided its customers with a direct route onto Usenet.

The arrival of these Internet neophytes saw early 'flame wars' breaking out online, strictures about 'net abuse' taking shape and a general resentment amongst long-established, 'professional' and elite Internet users.

It would be true to say that having an AOL address made you fair game for online abuse from intolerant old-timers or CompuServe users. That's just the way it was, as the process of democratising the Internet got under way.

Another Usenet trend emerged, first noted in about 1993, when cohorts of 'newbie' AOL users stumbled online every time a new University intake arrived each September.

Many of the 'net's original unwritten principles, including observing 'netiquette', reading the 'FAQ' before annoying people with questions, and the need to 'RTFM' were ignored by those who neither understood nor cared.

The fact was that the Internet genie had been let out of the bottle, and everybody now wanted to get in on the action.

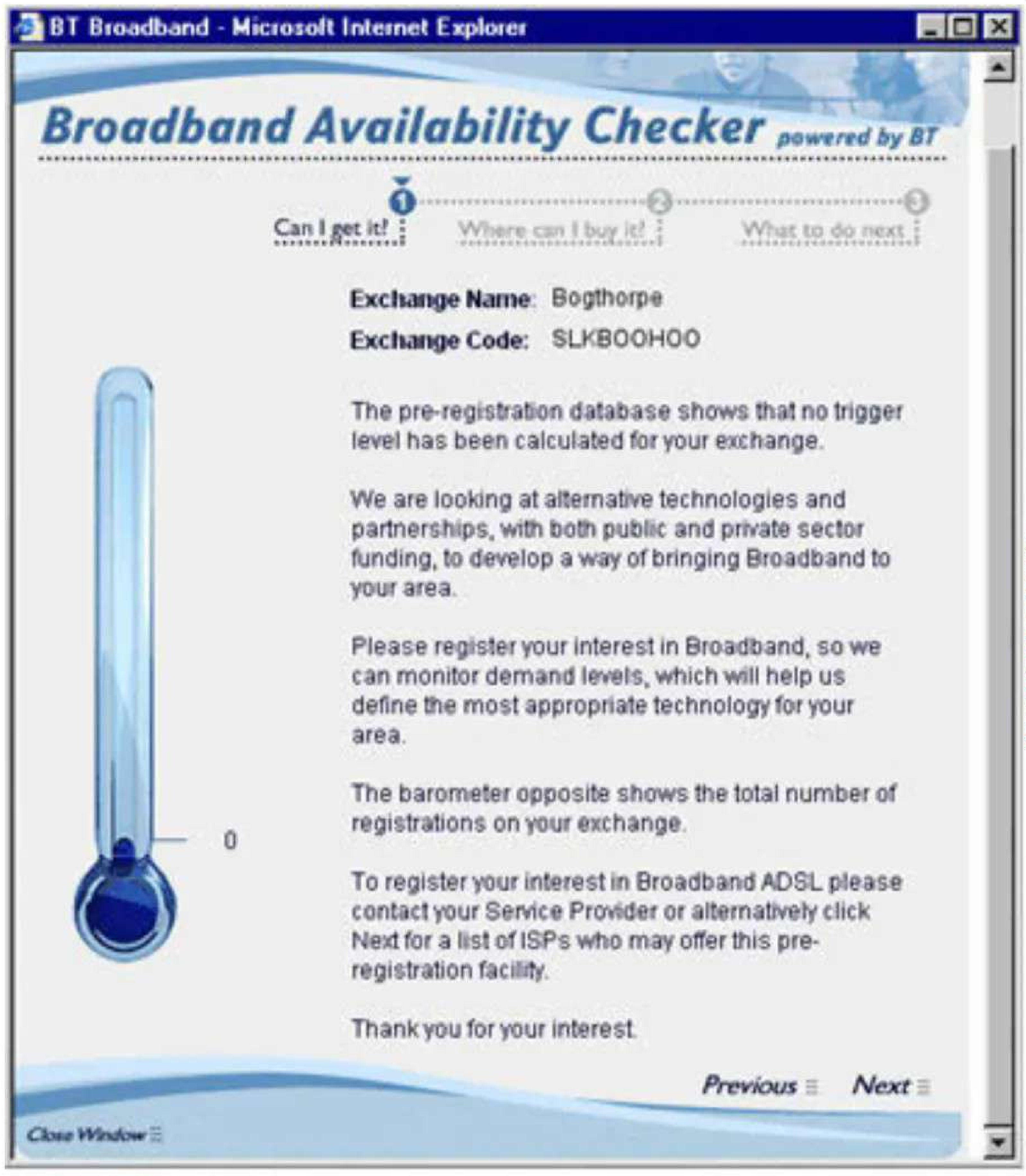
Portal websites, chat rooms, search engines and messaging systems were popularised as the Internet evolved at speed. Early forms of the eBay website described itself as "the world's online marketplace".

Numerous online brands such as Alta Vista, Ask Jeeves, Excite and ICQ came and went, while others, such as Yahoo! and Netscape, still hang on in there today. Ironically, AOL eventually acquired CompuServe.

The need for speed

By the early 2000s, new ADSL (Asymmetric Digital Subscriber Line) 'broadband' services promised maybe 6-10Mbps download speeds using existing copper phone lines. Initially, many localities, including my rural one, were ruled out from receiving ADSL, as the focus was on kitting out urban locations instead.

BT eventually introduced a system of drip-feeding the rollout of broad-


BT Broadband - Microsoft Internet Explorer
Broadband Availability Checker powered by BT
Can I get it?
Where can I buy it?
What to do next
0
Previous
Next
Close Window

Exchange Name: Bogthorpe
Exchange Code: SLKBOOHOO

The pre-registration database shows that no trigger level has been calculated for your exchange.

We are looking at alternative technologies and partnerships, with both public and private sector funding, to develop a way of bringing Broadband to your area.

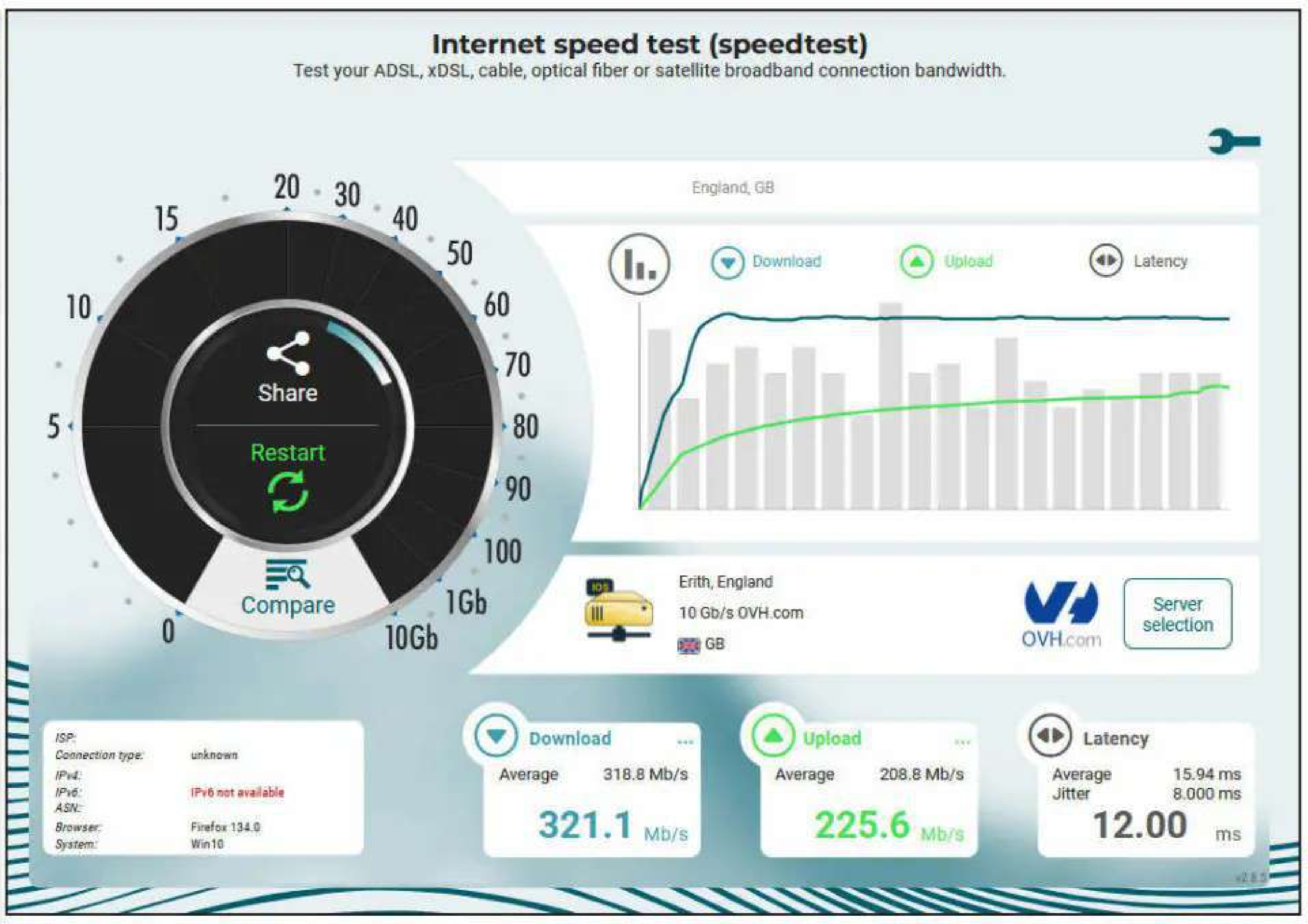
Please register your interest in Broadband, so we can monitor demand levels, which will help us define the most appropriate technology for your area.

The barometer opposite shows the total number of registrations on your exchange.

To register your interest in Broadband ADSL please contact your Service Provider or alternatively click Next for a list of ISPs who may offer this pre-registration facility.

Thank you for your interest.

BT's unloved Broadband Availability Checker displayed local 'expressions of interest' before BT would consider installing ADSL.


Internet speed test (speedtest)
Test your ADSL, xDSL, cable, optical fiber or satellite broadband connection bandwidth.
0
5
10
15
20
30
40
50
60
70
80
90
100
1Gb
10Gb
Share
Restart
Compare
England, GB
Download
Upload
Latency
OVH.com
Server selection

Erith, England
10 Gb/s OVH.com
GB

ISP:
Connection type: unknown
IPv4:
IPv6: IPv6 not available
ASN:
Browser: Firefox 134.0
System: Win10

Download
Average 318.8 Mb/s
321.1 Mb/s

Upload
Average 208.8 Mb/s
225.6 Mb/s

Latency
Average 15.94 ms
Jitter 8.000 ms
12.00 ms

The speed checker at **nPerf.com** will display your Internet connection's throughput and latency.

band by collating local 'expressions of interest'. A tantalising thermometer-style Broadband Availability Checker showed when local targets had been reached. Many users hated it, and they faced quite a cruel wait until ADSL broadband expanded in the mid-2000s.

BT then implored the trendy 'new media' industry to come up with multimedia video, E-commerce, chat applications and database-driven websites that could exploit these higher broadband speeds (more likely to fuel the demand for ADSL).

Consequently, websites became more bloated and over-designed, in sharp contrast to the tighter 'information silos' that had preceded them.

In this era, the so-called 'Web 2.0' was born, and email or surfing the web eventually became routine. Fibre-to-the-cabinet (FTTC) then promised faster-still data transmission rates, the 'last mile' being routed over existing copper wires to the property.

The Internet is now firmly established as the fourth essential utility, and several young generations have never known life without it and all the pressures that it brings. Now Wi-Fi, mobile smartphone data, satellite services and fibre-optic based data comms bring unlimited network access to everyday users, almost anywhere in the world.

Media is now streamed to devices via an Internet feed, and homes bristle with all manner of smart 'Internet of Things' devices connected to the cloud.

The world has become smaller, and it has been fascinating to see the impact that the Internet has had on our culture – quite a significant one, as it turns out.

Reaching the peak

My nostalgic journey over the past 30 years brings me full circle to the present day. My work lab, which started life with 14.4k dial-up, has reached what is probably 'peak Internet' as, against all odds, I successfully installed fibre to the property a few months ago.

Thanks to a local 'alt-net' supplier, I now enjoy a more than adequate 300Mbps download speed, with up to 1Gbps being possible – 70,000 times faster than my original dial-up! You can check your own broadband speed at **nPerf.com** – see the accompanying screenshot above.

My BT Openreach modem was replaced by an Optical Network Terminal (ONT), which terminates the optical fibre that comes in through an outside wall (upstairs in my case – no need to dig up the road).

My old SOHO router was swapped for a TP-Link EX230v, which has a separate VoIP (phone) port. Not everyone wants to be glued to their smartphone all day, but what I thought would be the trickiest part – migrating my BT landline number to the fibre service – went perfectly and it was transferred on time, costing just £5 a month extra.

Voice calls now have a better quality than I can ever remember. Note that factory-resetting a VoIP router will erase those landline config details, something that your ISP may need to restore again.

Mains supply outages are another consideration, so my compact 9V and 12V UPS made by Power Inspired (see *Net Work*, December 2024) will hopefully power the ONT and router for a reasonable period.

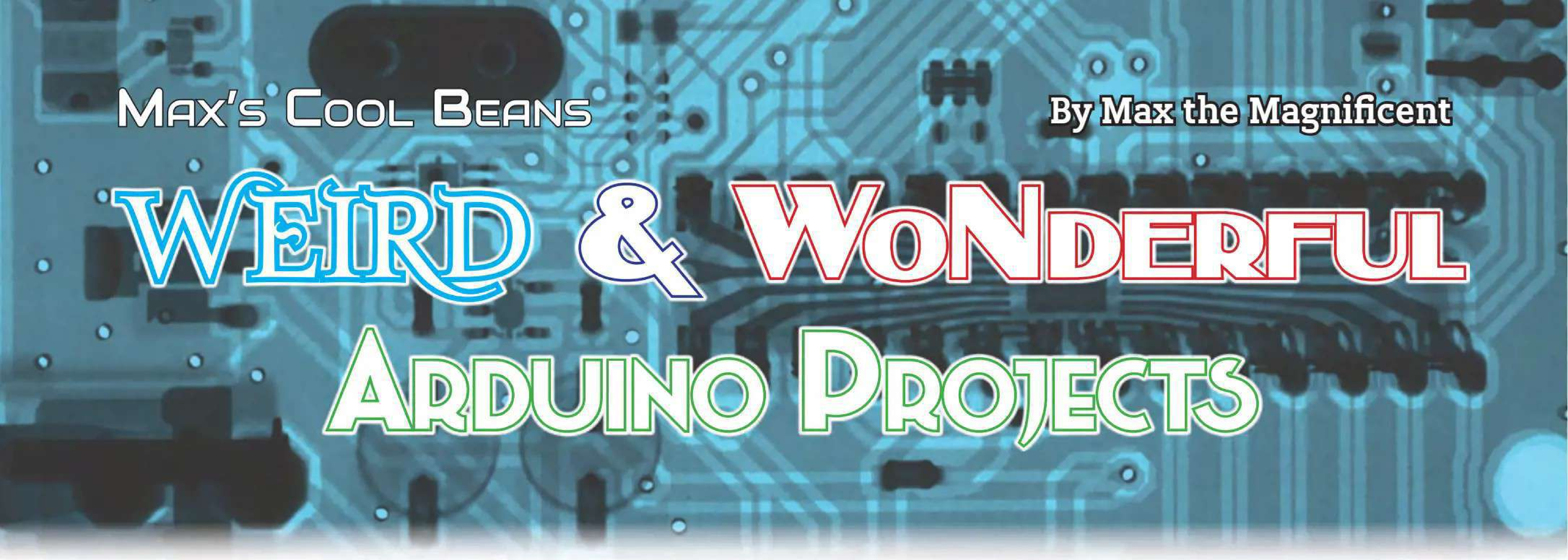
Installation and setup were free, thanks to the UK Government's Gigabit Broadband Voucher Scheme (GBVS). The information at **https://pemag.au/link/ac4j** was incorrect, though, and you should check the score with your chosen fibre supplier first.

Pulling the plug

I'm told that the entire world's phone calls could pass through my single fibre optic connection. Maybe that's sales patter, but things have come a long way since the 1990s when, just to check my email, I would unfurl a phone extension cable across the landing, like a sapper laying explosives, so that I could plug my modem into the phone socket upstairs.

Doubtless, the arrival of full fibre broadband will open up new avenues to explore and keep me entertained and interested. With this in mind, I must also share the news that this month's column is the last in the series of regular *Net Work* articles. After nearly 29 years, I'm now taking a well-earned break, but *Net Work* contributions may still appear periodically well into the future.

I hope they've helped inform readers of new trends heading our way, as well as offering handy 'heads-up' reminders about getting the best from some current technologies. Many thanks to my regular readers for your feedback and kind comments and, in the meantime, I can still be contacted by email to **alan@epemag.net** as always, or via my website at **www.alanwinstanley.com** PE



Max's Cool Beans

By Max the Magnificent

Weird & Wonderful Arduino Projects

Part 4: full power ahead (to the main array)!

My head hurts; I'm about to make your head hurt too (you're welcome.) As you may recall, in our previous column, we used our strip of 144 tricolour light-emitting diodes (LEDs) to create the 14 × 10 (width × height) array that will form the main display area for our retro games console. This left us with four unused LEDs; fear not, we'll find a use for these later (so don't lose them).

We decided to refer to each of the LEDs in our array as 'pixels' (short for picture elements), each of which has red, green, and blue (RGB) subelements. We also decided to view our array in terms of (**x**, **y**) coordinates and to employ a 'bottom-left' origin system, in which (**x**, **y**) = (0, 0) is in the lower-left corner, and (**x**, **y**) = (13, 9) is in the upper-right corner.

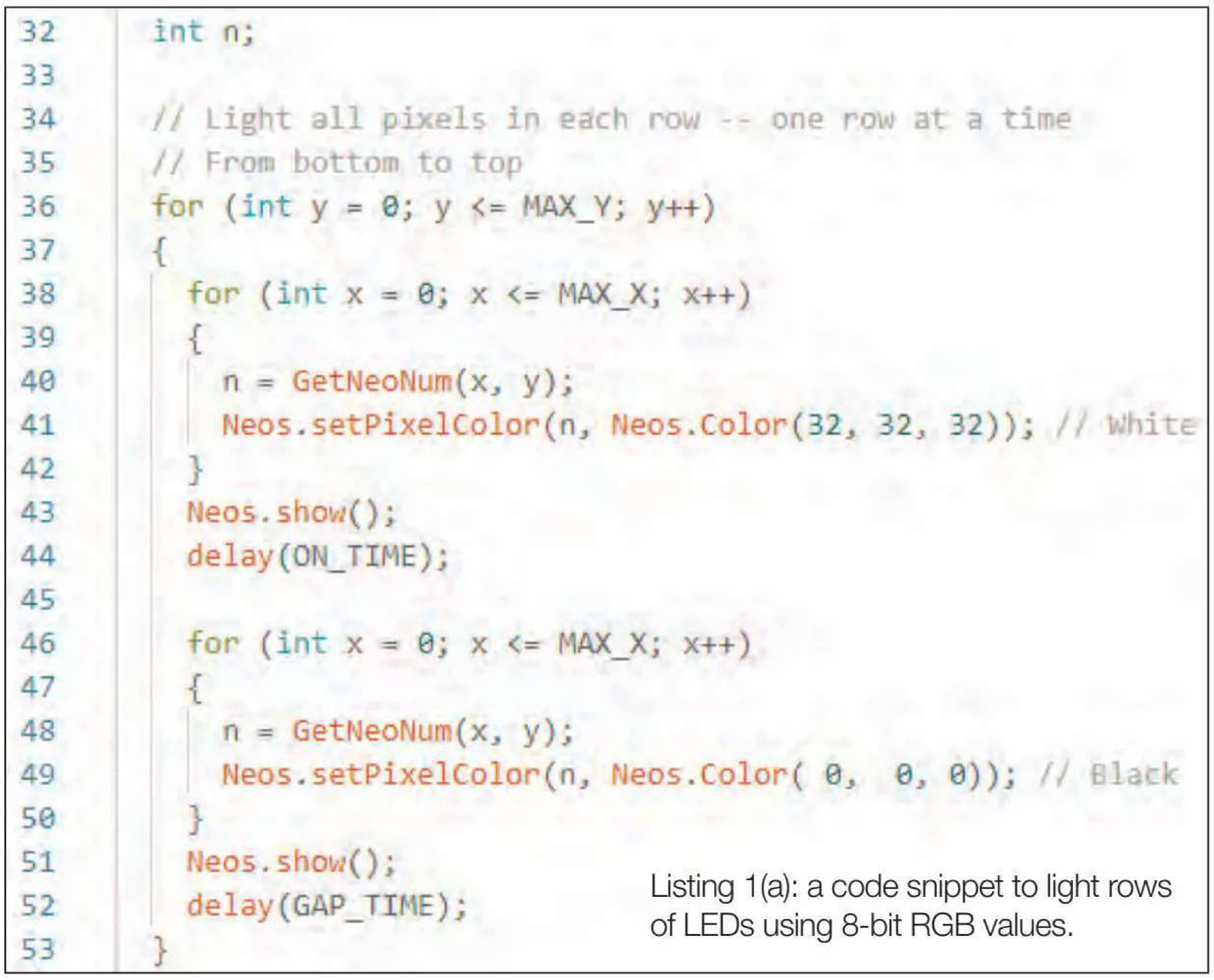
We noted that it's possible to connect our ten 14-pixel segments in a variety of different ways and to sort things out in software later. To this end, once we'd verified that everything worked, we created a **GetNeoNum()** function that accepts **x** and **y** integer values as arguments and returns the number **n** of the corresponding pixel in the string.

As I noted in that column, the great thing about this is that it doesn't matter if you organise your segments differently to me. So long as we both have **GetNeoNum()** functions that work with our own arrays, and with identical external interfaces, all you need to do to run one of my programs is to exchange your version of this function for mine.

After performing various experiments, we ended by creating a program that repeatedly (a) lit each row from bottom to top, (b) lit each row from top to bottom, (c) lit each column from left to right, and (d) lit each column from right to left. Since we are currently powering our array from the Arduino, we made sure that only a single row or column was active at any time.

To refresh your memory, you can download a copy of this code in the file named **CB-Apr25 Code-01.txt**. As usual, all the files mentioned in this column are available from the April 2025 page of the PE website: **https://pemag.au/link/ac44**

I left you pondering how we could extend our program to light the pixels in diagonal lines, one line at a time, starting with the one-pixel diagonal 'line' in the bottom left corner at (**x**, **y**) = (0, 0) and ending with the 1-pixel diagonal 'line' in the top right corner at (**x**, **y**) = (13, 9). This is what we are poised do shortly but we'll take things step-by-step to ensure we are all dancing to the same tune.



```
32      int n;
33
34      // Light all pixels in each row -- one row at a time
35      // From bottom to top
36      for (int y = 0; y <= MAX_Y; y++)
37      {
38        for (int x = 0; x <= MAX_X; x++)
39        {
40          n = GetNeoNum(x, y);
41          Neos.setPixelColor(n, Neos.Color(32, 32, 32)); // White
42        }
43        Neos.show();
44        delay(ON_TIME);
45
46        for (int x = 0; x <= MAX_X; x++)
47        {
48          n = GetNeoNum(x, y);
49          Neos.setPixelColor(n, Neos.Color( 0,  0, 0)); // Black
50        }
51        Neos.show();
52        delay(GAP_TIME);
53      }
```

Listing 1(a): a code snippet to light rows of LEDs using 8-bit RGB values.

Black and white

Let's start by looking at the portion of our existing program that lights each row from bottom to top as illustrated in Listing 1(a). The listing number (1 in this case) corresponds to part of the name of the matching code file ("01" here).

Remember that **Neos** is the identifier we've chosen for our pixel strings. Observe the calls to the **Neos.setPixelColor()** function on lines 41 and 49. This function accepts two arguments. The first is the number of the pixel whose colour we wish to change. The second is a 24-bit colour value comprising 8 bits each for the RGB components.

Also remember that this function only modifies the colour of the pixel stored in the Arduino's memory. Nothing changes in the outside world until we call the **Neos.show()** function on lines 43 and 51 to stream the values from memory into the physical string of pixels forming our array.

Currently, we are generating these 24-bit values using the **Neos.Color()** function, which accepts three 8-bit RGB values and combines them into a single 24-bit value.

Our compiler understands only 8-bit, 16-bit, and 32-bit integer values (most computers have no native support for three-byte [24-bit] integers). As a result, our 24-bit colours are stored in 32-bit variables and we simply ignore the most-significant (uppermost) eight bits, which remain unused.

Each of our 8-bit RGB fields can have values ranging from 0 to 255 in decimal, which equates to 0b00000000 to 0b11111111 in binary (the '0b' indicates binary) or 0x00 to 0xFF in hexadecimal (the '0x' indicates hexadecimal). Driving the RGB fields with all the same values results in white light.

The reason we are using RGB values of 32 (0x20) on line 41 is to reduce the amount of power we are consuming to 32 ÷ 255 ≃ 12.5% of full power.

The data sheet tells us that, when fully on, each of our pixels consumes 60mA (milliamperes); that's 20mA for each of the RGB components. So, when displaying white light at 12.5% of full power, each pixel consumes 60mA × 0.125 = 7.5mA. This calculation assumes everything is linear, which may not be the case, but this approximation is 'close enough for government work', as they say.

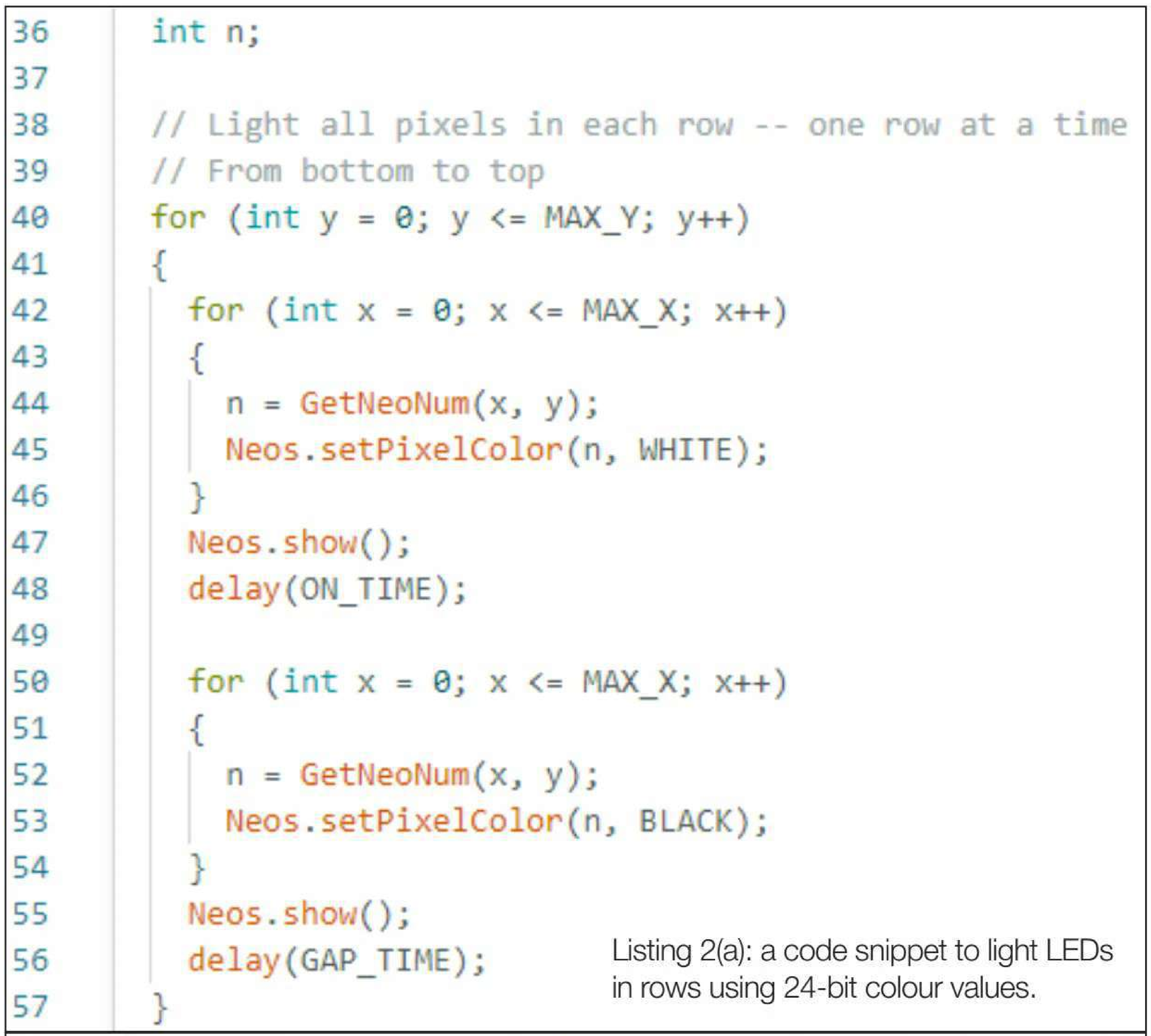
In the case of our existing program, our worst case is for all 14 pixels on the same row to be lit up at the same time. The resulting 7.5mA × 14 = 105mA is well within the capabilities of our Arduino.

As opposed to using the **Neos.Color()** function, we can simply define our colours as 24-bit values:

```
#define WHITE 0x202020
#define BLACK 0x000000
```

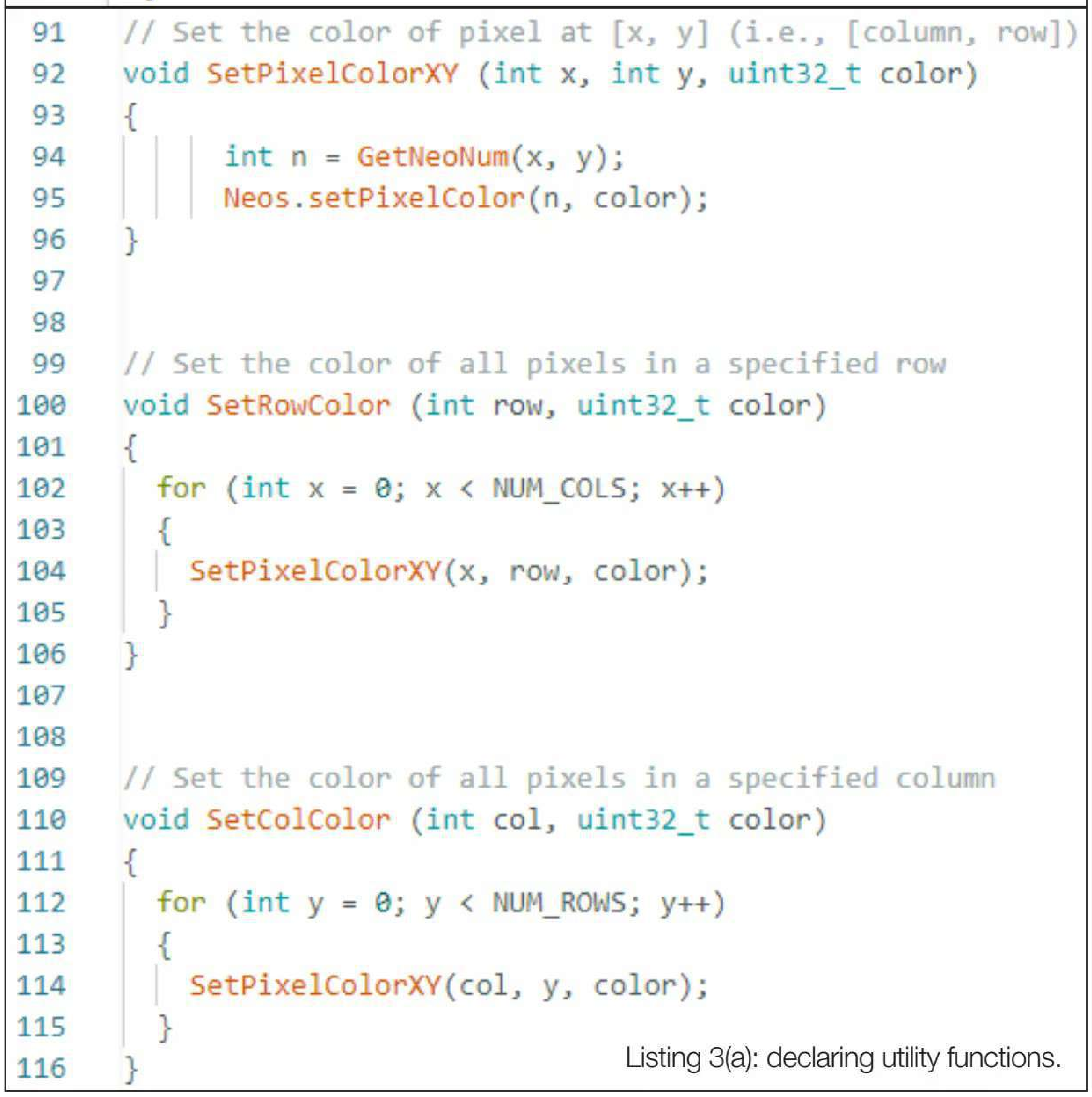
This allows us to modify (and simplify) the calls to our **Neos.setPixelColor()** functions, as seen in lines 45 and 53 of Listing 2(a). You can download the new version of our program in the file named **CB-Apr25 Code-02.txt**.

Although this may not appear to be a great leap forward, working with 24-bit colour values will make our lives a lot easier in the future.



```
36    int n;
37
38    // Light all pixels in each row -- one row at a time
39    // From bottom to top
40    for (int y = 0; y <= MAX_Y; y++)
41    {
42      for (int x = 0; x <= MAX_X; x++)
43      {
44        n = GetNeoNum(x, y);
45        Neos.setPixelColor(n, WHITE);
46      }
47      Neos.show();
48      delay(ON_TIME);
49
50      for (int x = 0; x <= MAX_X; x++)
51      {
52        n = GetNeoNum(x, y);
53        Neos.setPixelColor(n, BLACK);
54      }
55      Neos.show();
56      delay(GAP_TIME);
57    }
```

Listing 2(a): a code snippet to light LEDs in rows using 24-bit colour values.



```
 91  // Set the color of pixel at [x, y] (i.e., [column, row])
 92  void SetPixelColorXY (int x, int y, uint32_t color)
 93  {
 94      int n = GetNeoNum(x, y);
 95      Neos.setPixelColor(n, color);
 96  }
 97
 98
 99  // Set the color of all pixels in a specified row
100  void SetRowColor (int row, uint32_t color)
101  {
102    for (int x = 0; x < NUM_COLS; x++)
103    {
104      SetPixelColorXY(x, row, color);
105    }
106  }
107
108
109  // Set the color of all pixels in a specified column
110  void SetColColor (int col, uint32_t color)
111  {
112    for (int y = 0; y < NUM_ROWS; y++)
113    {
114      SetPixelColorXY(col, y, color);
115    }
116  }
```

Listing 3(a): declaring utility functions.

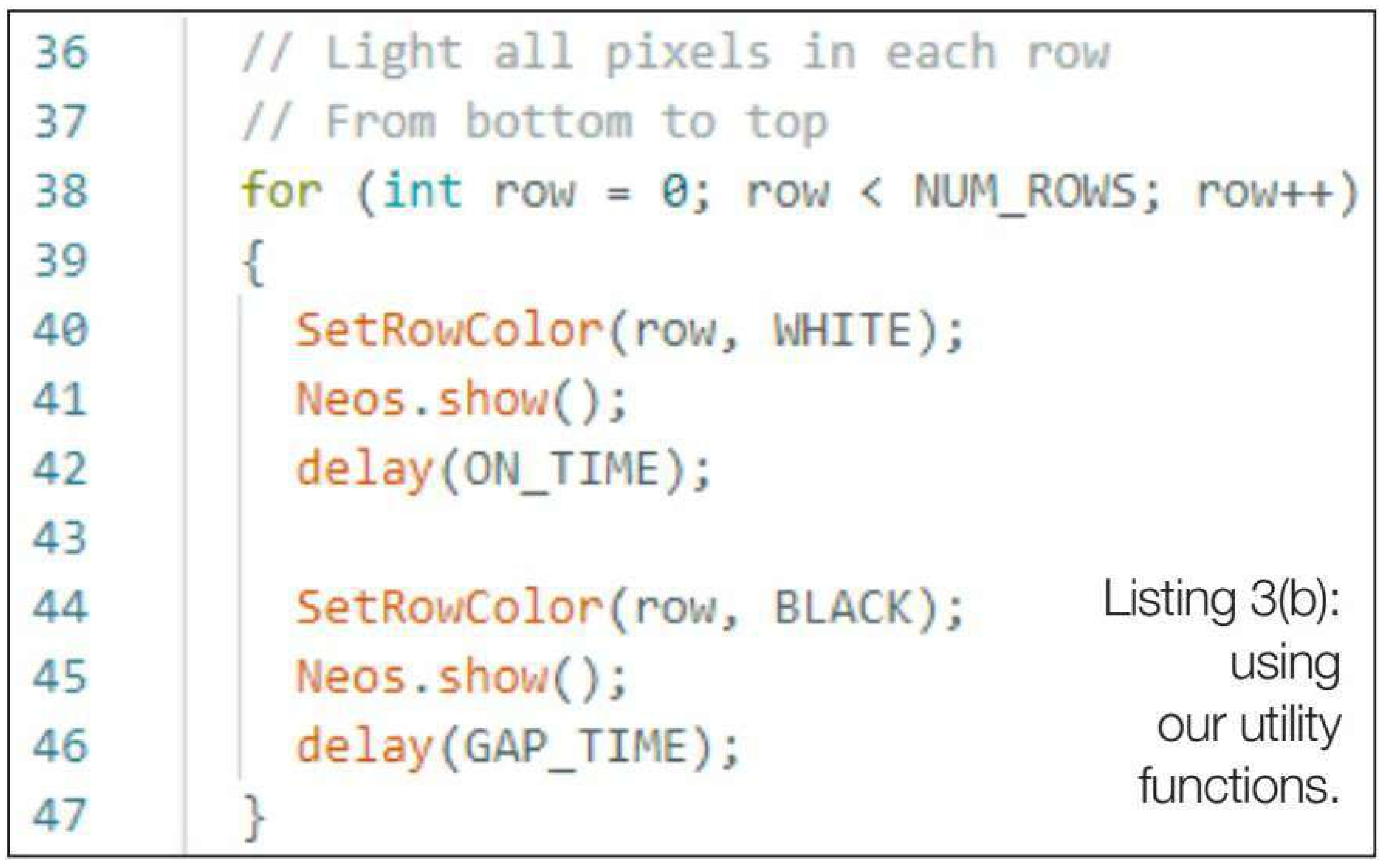
Declaring utility functions

Most of the code in the main body of our program currently lives in the **loop()** function. Instead of writing things as one large 'blob' of code (I hope I'm not getting too technical), a fundamental practice in programming is to create a suite of low-level utility functions. These functions can then be called from other parts of the program (they can even call each other).

Employing utility functions improves the readability, reusability, and scalability of our code. It also makes our programs easier to debug and maintain. What's not to love?

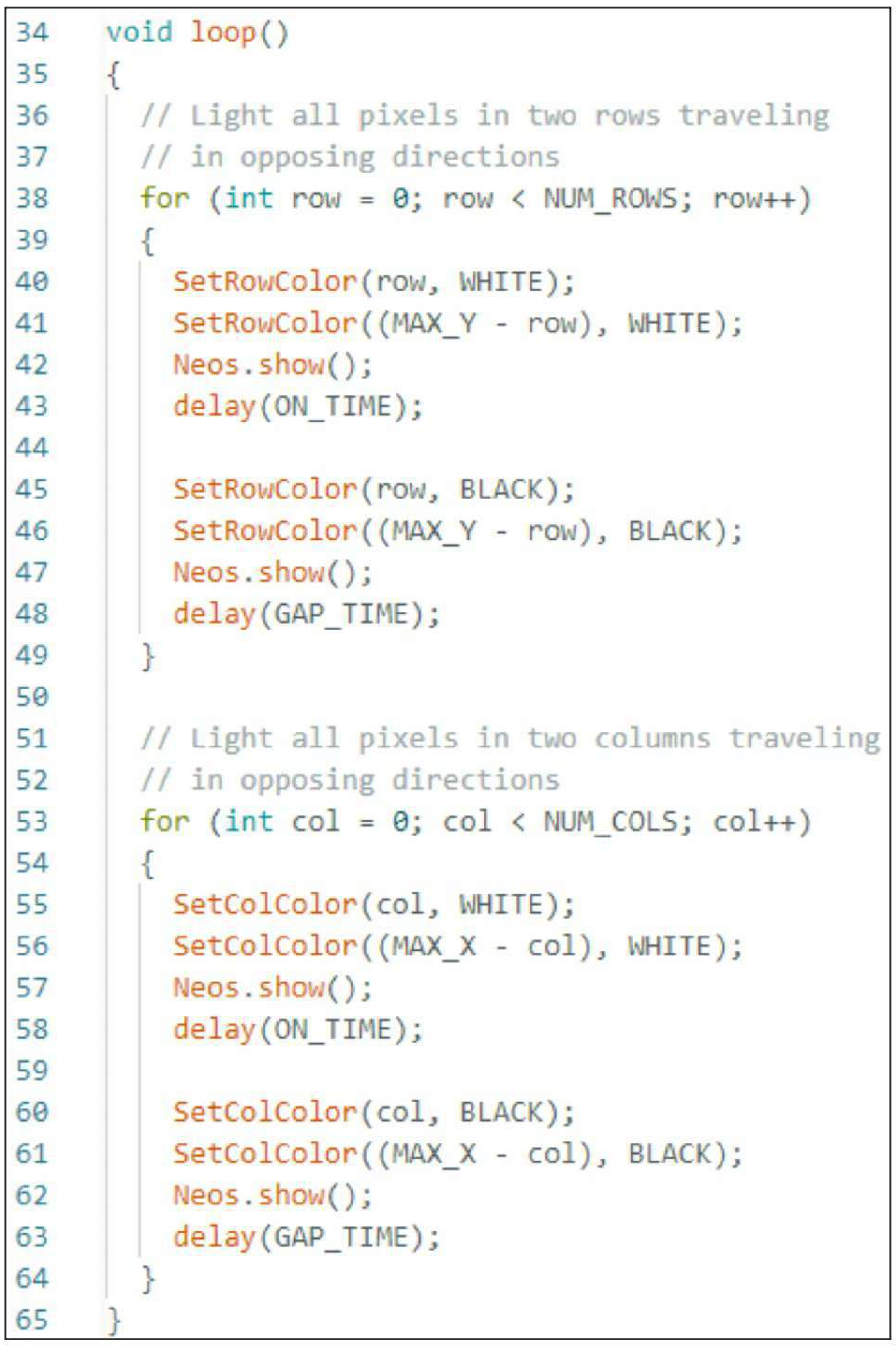
We already have one utility function in the form of our **GetNeoNum()** routine. The next iteration of our program (in the file named **CB-Apr25 Code-03.txt**) contains three more, as seen in Listing 3(a).

The **SetPixelColorXY()** function at line 92 sets the colour of a single pixel (once again, it modifies the pixel value stored in the Arduino's memory, not in the physical string). This function accepts three parameters: the **x** and **y** coordinates of the pixel in question, and



```
36    // Light all pixels in each row
37    // From bottom to top
38    for (int row = 0; row < NUM_ROWS; row++)
39    {
40      SetRowColor(row, WHITE);
41      Neos.show();
42      delay(ON_TIME);
43
44      SetRowColor(row, BLACK);
45      Neos.show();
46      delay(GAP_TIME);
47    }
```

Listing 3(b): using our utility functions.



```
34  void loop()
35  {
36    // Light all pixels in two rows traveling
37    // in opposing directions
38    for (int row = 0; row < NUM_ROWS; row++)
39    {
40      SetRowColor(row, WHITE);
41      SetRowColor((MAX_Y - row), WHITE);
42      Neos.show();
43      delay(ON_TIME);
44
45      SetRowColor(row, BLACK);
46      SetRowColor((MAX_Y - row), BLACK);
47      Neos.show();
48      delay(GAP_TIME);
49    }
50
51    // Light all pixels in two columns traveling
52    // in opposing directions
53    for (int col = 0; col < NUM_COLS; col++)
54    {
55      SetColColor(col, WHITE);
56      SetColColor((MAX_X - col), WHITE);
57      Neos.show();
58      delay(ON_TIME);
59
60      SetColColor(col, BLACK);
61      SetColColor((MAX_X - col), BLACK);
62      Neos.show();
63      delay(GAP_TIME);
64    }
65  }
```

Listing 4(a): controlling multiple rows and columns of LEDs.

the **color** we wish to assign to that pixel.

The **SetRowColor()** function at line 100 sets all the pixels in a row to the same colour. It accepts two parameters: the **row** in question (from 0 to 9) and the **color** we wish to assign to the pixels in that row.

Similarly, the **SetColColor()** function at line 110 sets all the pixels in a column to the same colour. It accepts two parameters: the **col** in question (from 0 to 13), and the **color** we wish to assign to the pixels in that column.

Observe that, in all three functions, we've declared the colour variable as being of type **uint32_t**, which refers to a 32-bit (4-byte) unsigned (non-negative) integer. Ideally, we would prefer to use a **uint24_t**, but this type isn't supported by most compilers.

Happily, when we pass a 24-bit value into our function, the compiler will automatically pad the unused eight most-significant bits with 0s to generate the 32-bit value expected by the function.

You may be wondering why we used a data type of **uint32_t** instead of an **unsigned long**, which is what we would expect to see in most Arduino documentation. We don't want to get into this here, but we will discuss the various integer data types in ~~excruciating~~ exhilarating detail in next month's column.

Observe that our **SetPixelColorXY()** function calls our existing **GetNeoNum()** function. Similarly, both our **SetRowColor()** and **SetColColor()** functions call our **SetPixelColorXY()** function. The result is a cascading hierarchy of function calls.

Using our utility functions

Now that we've declared our utility functions, we can modify our **loop()** function to take advantage of them. Let's consider the new incarnation of the portion of our program that lights each row from bottom to top as illustrated in Listing 3(b).

As is always the case with programming, there are many different approaches we could use. For example, in our **SetRowColor()** function, we could move the **show()** and **delay()** function calls on lines 41 and 42 (and on lines 45 and 46) into the body of the **SetRowColor()** function itself.

Since we use different delay values at different times, we could pass the delay duration into the **SetRowColor()** function as an additional argument.

As tempting as this may be, it would place limitations on how we could use our **SetRowColor()** function for other tasks in the future. For example, suppose we decided that we wished to modify our **loop()** function to start by lighting rows 0 and 9 (bottom and top), then turn these rows off and light rows 1 and 8, then 2 and 7, then 3 and 6, ending with 9 and 0.

This is easy-peasy lemon squeezy with our existing **SetRowColor()** function, but we would be stressed-depressed lemon zest if we had modified our function as discussed above.

To illustrate this in action. I created a dual-row version of the program using our existing **SetRowColor()** function (in the file named **CB-Apr25 Code-04.txt**).

As illustrated in Listing 4(a), the modified manipulation of our rows occurs from line 38 to line 49. I've also implemented a similar treatment to work with dual columns, as seen from line 53 to line 64.

Since we are now lighting two rows (or two columns) at the same time, our new worst-case current draw will be 210mA, which is still well within the capabilities of our Arduino.

Observe the way in which we set two rows to be **WHITE** on lines 40 and 41 before calling the **show()** function on line 42 and the **delay()** function on line 43. Similarly, we set the two rows to be **BLACK** on lines 45 and 46 before calling the **show()** function on line 47 and the **delay()** function on line 48.

The point of all this is that we are performing a programming balancing act. In one possible scenario, we don't create any utility functions, but instead implement everything as a great big glob of code in the main **loop()** function.

At the other end of the spectrum, we stick everything into separate functions, but this could make them clunky and unwieldy, possibly requiring us to create multiple versions of each function to satisfy the needs of different use cases.

The 'Goldilocks' solution is to put just the right amount of code into our utility functions (not too much and not too little). The trick is knowing where to draw the metaphorical line.

Dutch angles

This is where we return to the task of lighting our pixels in diagonal lines across the display, one line at a time. We will start with what I think of as 'backward slash diagonals' (like the \ character). That is, starting with the one-pixel 'line' in the bottom left corner

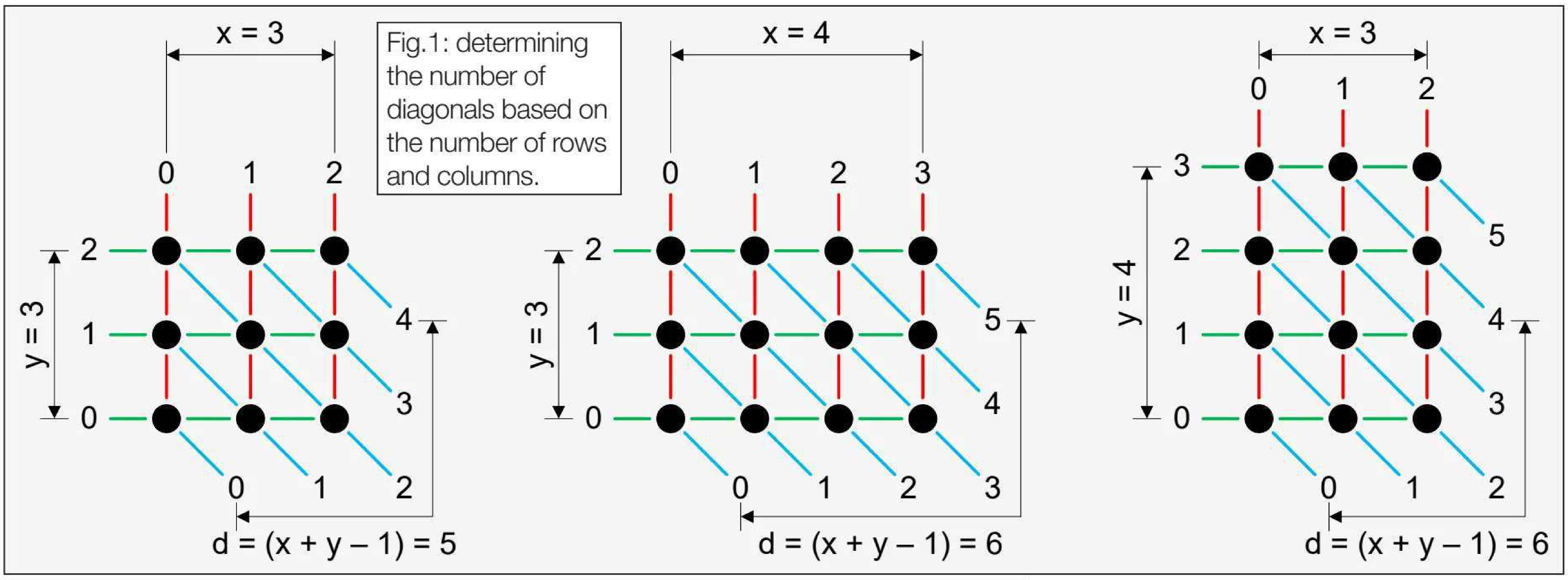


Fig.1: determining the number of diagonals based on the number of rows and columns.

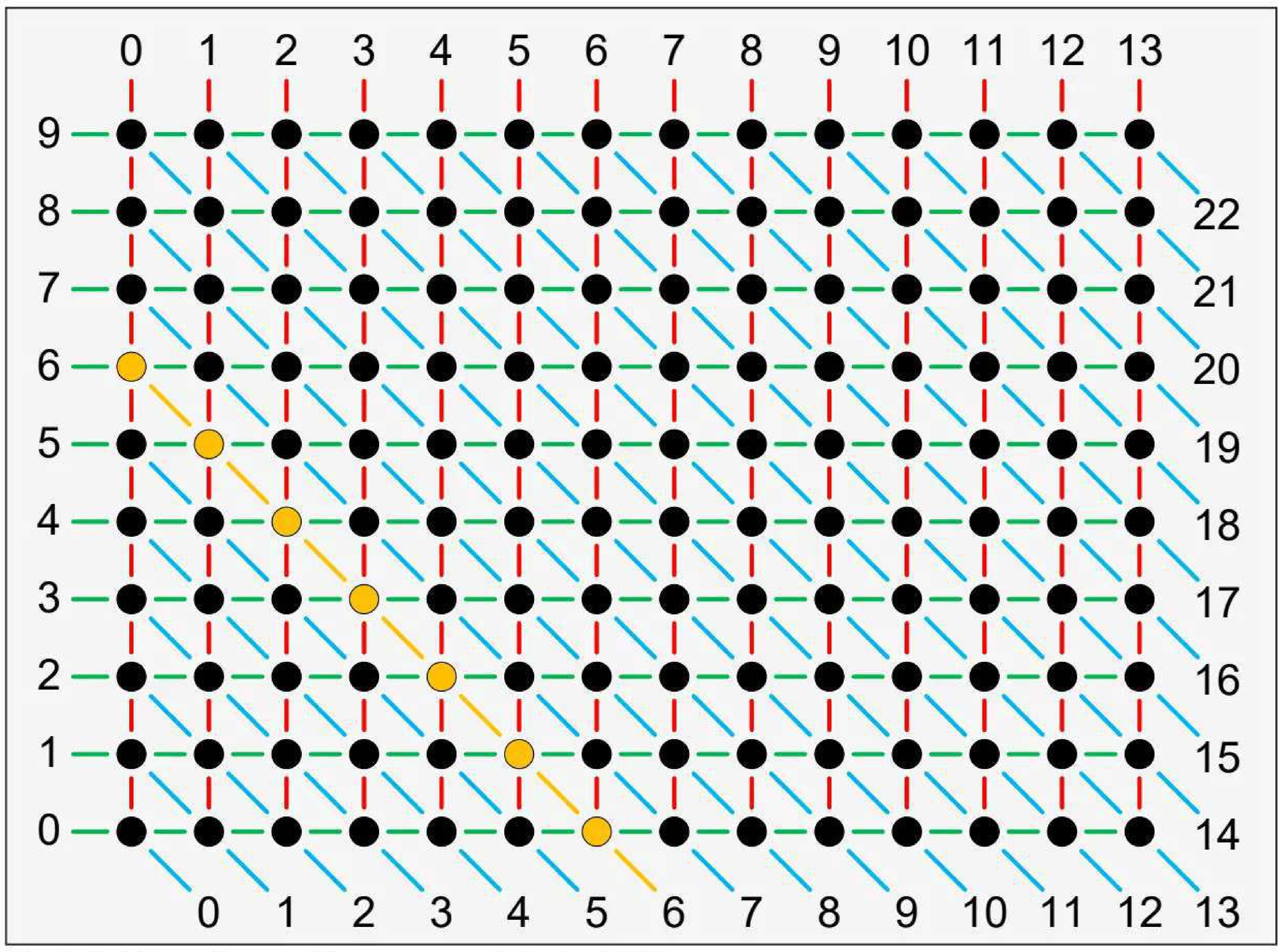


Fig.2: lighting diagonal lines on our full-size LED array.

at (**x**, **y**) = (0, 0) and ending with the 1-pixel 'line' in the top right corner at (**x**, **y**) = (13, 9).

Once we have this working, your homework assignment this month will be to implement what I think of as 'forward slash diagonals' (like the / character). That is, starting with the 1-pixel 'line' in the top left corner at (**x**, **y**) = (0, 9) and ending with the 1-pixel 'line' in the bottom right corner at (**x**, **y**) = (13, 0).

Before we proceed, for any maths nerds amongst us, I should admit to being a little loose with my use of the word "diagonal". In mathematics, a diagonal is a line that connects two vertices of a polygon or a solid where these vertices are not on the same edge.

In the case of a square or rectangle, for example, we are talking about the line that connects the top left and bottom right corners or the line that connects the bottom left and top right corners.

In our case, by comparison, I'm using 'diagonal' to mean a line 45° from horizontal. Actually, even this isn't true because our pixels aren't spaced equally in the horizontal and vertical dimensions. But let's not worry about that too much.

In the case of 'backward slash diagonals', diagonal 0 would be at pixel (**x**, **y**) = (0, 0); diagonal 1 would follow the (**x**, **y**) pixel path (0, 1), (1, 0); diagonal 2 would follow the (**x**, **y**) path (0, 2), (1, 1), (2, 0); diagonal 3 would follow the (**x**, **y**) path (0, 3), (1, 2), (2, 1), (3, 0); and so forth.

Our array has 10 horizontal rows and 14 vertical columns, so how many of our 45° diagonals do we have? As I've mentioned on previous occasions, I often find that sketching pictorial representations gives me insights that aid in my program creation activities. For example, consider the three small arrays shown in Fig.1.

From this we can deduce that, irrespective of the size of the array, the number of diagonals is one less than the sum of the number of rows and number of columns. In the case of our 14 × 10 array, this means we will have (14 + 10 – 1) = 23 diagonals.

As you may recall, our existing programs currently have four definitions associated with the number of rows and columns in our array. To these we can now add a definition associated with the number of diagonals, resulting in the following:

```
#define NUM_ROWS  10
#define NUM_COLS  14
#define NUM_DIAGS 23
#define MAX_X     13
#define MAX_Y      9
```

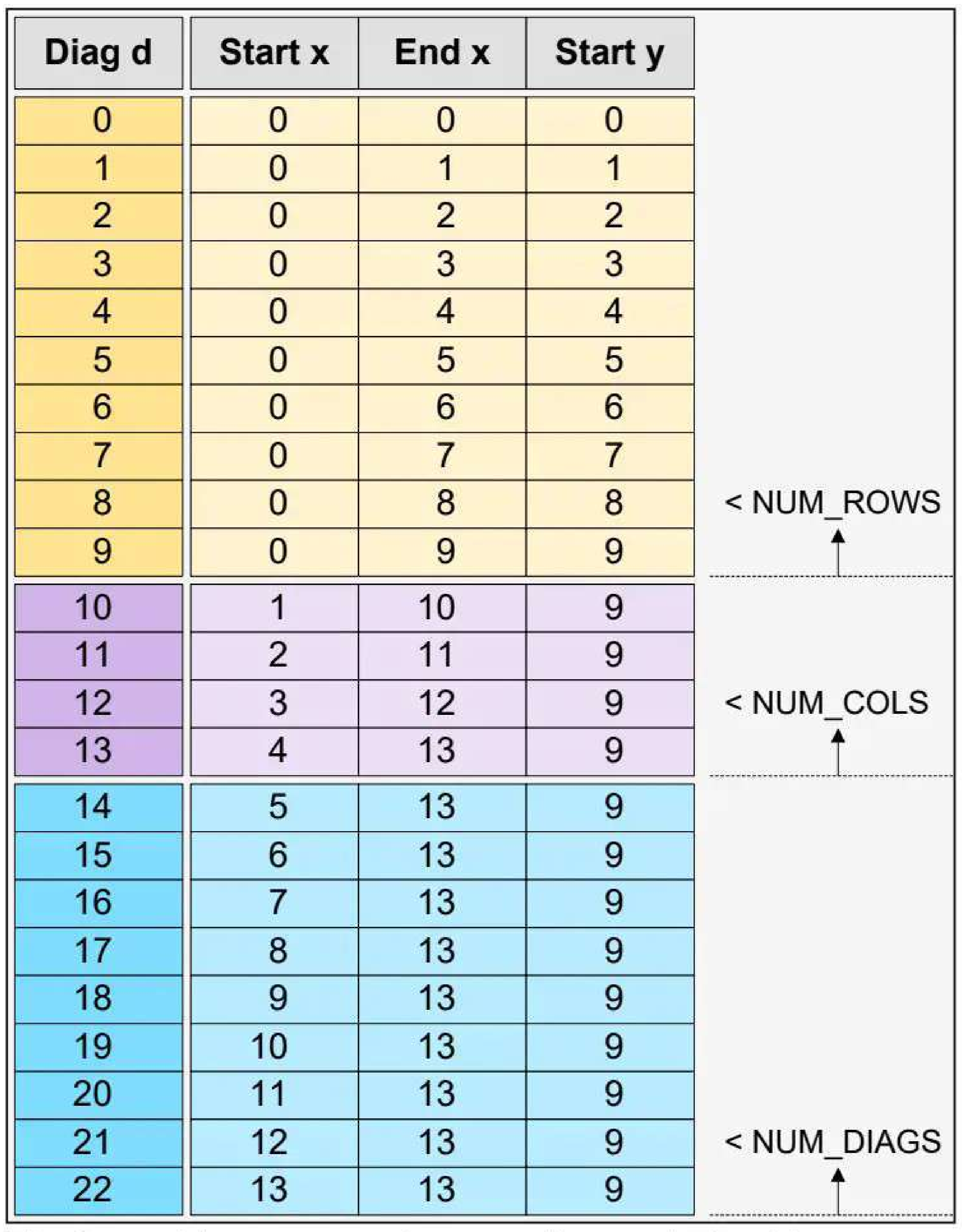
Instead of 23, if we wanted to be really clever, we could put **(NUM_ROWS + NUM_COLS – 1)**. Then we wouldn't need to change the value if we changed the number of rows or columns.

Now let's scale things up to our full-size array (Fig.2). In the real world, since we can perform only one task at a time, when we 'draw' our backward slash diagonals on the display, we must start at one end and work our way to the other. There are two ways to do this: either we start with the upper-left pixel and end on the lower-right pixel (the approach I opted for), or vice versa.

So, what algorithm are we going to use? The way I usually approach this sort of thing is to pick an element at random and see if I can deduce anything from it. We'll start with diagonal **d** = 6 (highlighted in orange) because 6 is my lucky number.

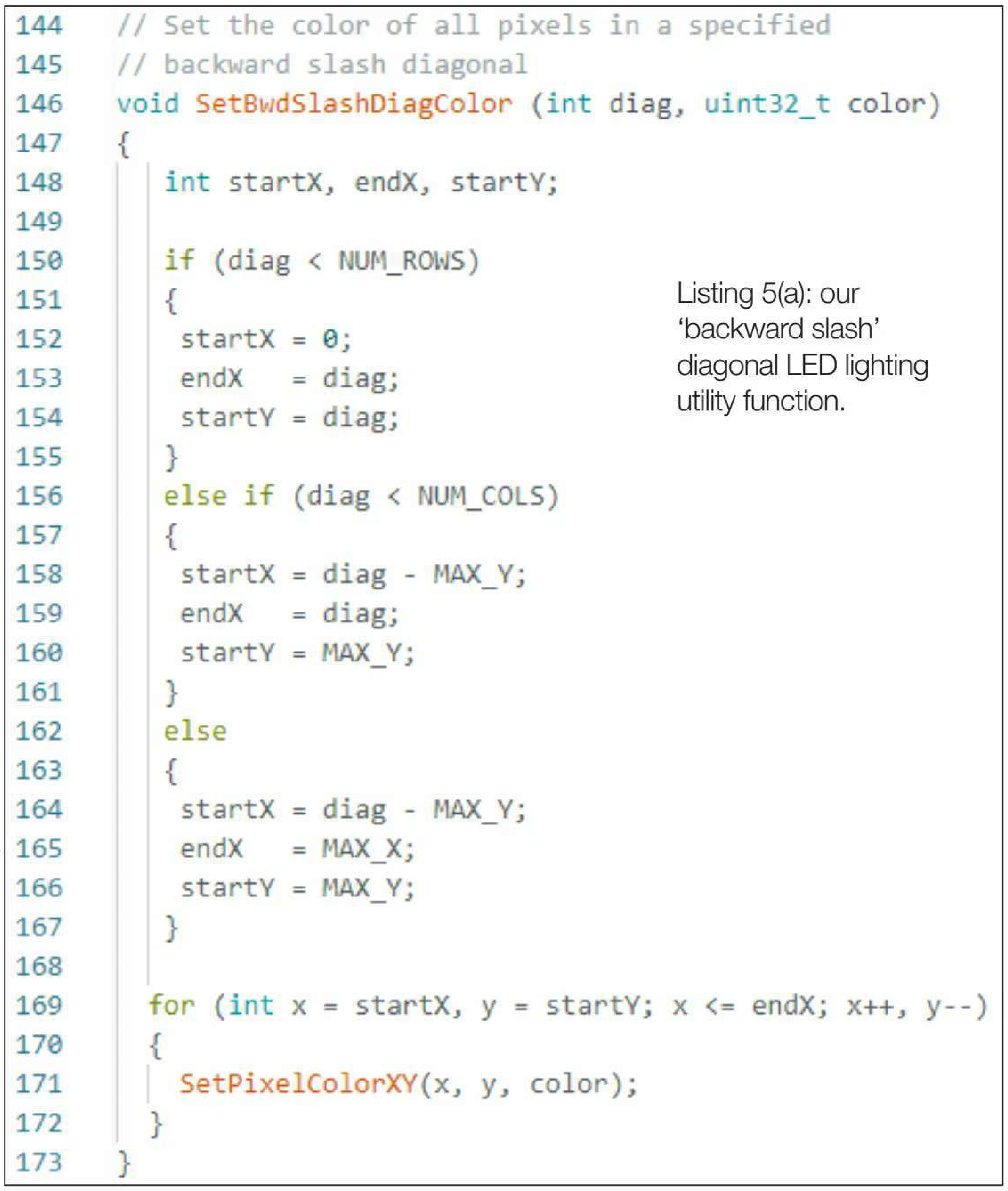
In this case, our start **x** value will be 0, our end **x** value will be 6, and our start **y** value will also be 6. Hmmm, this means that end **x** and start **y** are both equal to **d**. Well, that's convenient, but when we delve a little deeper, we discover that it doesn't work for all the diagonals.

Another tool I often use to provide insights for my program creation activities is to create tabular representations of things. For example,



Diag d	Start x	End x	Start y	
0	0	0	0	
1	0	1	1	
2	0	2	2	
3	0	3	3	
4	0	4	4	
5	0	5	5	
6	0	6	6	
7	0	7	7	
8	0	8	8	< NUM_ROWS
9	0	9	9	
10	1	10	9	
11	2	11	9	
12	3	12	9	< NUM_COLS
13	4	13	9	
14	5	13	9	
15	6	13	9	
16	7	13	9	
17	8	13	9	
18	9	13	9	
19	10	13	9	
20	11	13	9	
21	12	13	9	< NUM_DIAGS
22	13	13	9	

Fig.3: a table representing our diagonal algorithm.



```
144 // Set the color of all pixels in a specified
145 // backward slash diagonal
146 void SetBwdSlashDiagColor (int diag, uint32_t color)
147 {
148    int startX, endX, startY;
149
150    if (diag < NUM_ROWS)
151    {
152     startX = 0;
153     endX   = diag;
154     startY = diag;
155    }
156    else if (diag < NUM_COLS)
157    {
158     startX = diag - MAX_Y;
159     endX   = diag;
160     startY = MAX_Y;
161    }
162    else
163    {
164     startX = diag - MAX_Y;
165     endX   = MAX_X;
166     startY = MAX_Y;
167    }
168
169   for (int x = startX, y = startY; x <= endX; x++, y--)
170   {
171     SetPixelColorXY(x, y, color);
172   }
173 }
```

Listing 5(a): our 'backward slash' diagonal LED lighting utility function.

based on the full-size array in Fig.2, I created the table shown in Fig.3.

Three distinct areas jump out at us from the table shown in Fig 3. These areas are associated with the diagonal **d** being (a) less than **NUM_ROWS**, (b) greater than or equal to **NUM_ROWS** and less than **NUM_COLS**, or (c) greater than or equal to **NUM_COLS** and less than **NUM_DIAGS**.

This new intelligence allows us to create an updated representation of our array as illustrated in Fig.4. More importantly, the algorithm we are going to use practically jumps out of our tabular representation. Still, before we go there…

Foxy for() loops

As I mentioned many yonks ago (*PE*, May 2020), there's something rather clever about **for()** loops in C/C++ that professional programmers don't even think to mention, which may explain why many non-professional programmers don't know anything about it. Consider the following:

for(initialisation; condition; modification)

As we see, there are three semicolon-separated parts to this statement: **initialisation**, which is where we initialise our control variable; **condition**, which is where we test our control variable; and **modification**, which is where we modify our control variable. For example, we wouldn't be surprised to see a statement looking something like the following:

for (int i = 0; i < 10; i++)

There are several aspects to this that aren't immediately obvious to beginners.

Firstly, the modification isn't obliged to be a simple increment (++) or decrement (--), but can instead be a more complex expression, such as "i = i + 3".

Secondly, the condition can be as complicated as required, such as "((i < 10) && (j == 0))", for example.

Thirdly, both the initialisation and modification sections can comprise multiple comma-separated elements. We could have something like "i = 0, j = 0" in the initialisation section. Similarly, we could have something like "i++, j--" in the modification section. Let's keep all this in mind as we proceed to our…

Awesome algorithm

I created a new version of our program in the file named **CB-Apr25 Code-05.txt**. The first thing I did was to create an additional utility function called **SetBwdSlashDiagColor()**, illustrated in Listing 5(a). This sets all the pixels in a backward slash diagonal to the same colour.

It accepts two parameters: the **diag** ('diagonal') in question, from 0 to 22, and the **color** we wish to assign to the pixels in that diagonal.

Observe that on line 148, we declare three local integer variables: **startX**, **endX**, and **startY**. These reflect the columns in our table from Fig.3. We set the values of these variables based on the number of the diagonal we're working with. This occurs on lines 150 to 167. I dare to hope that the way this works is self-explanatory—just compare this portion of the code to the three sections of our table.

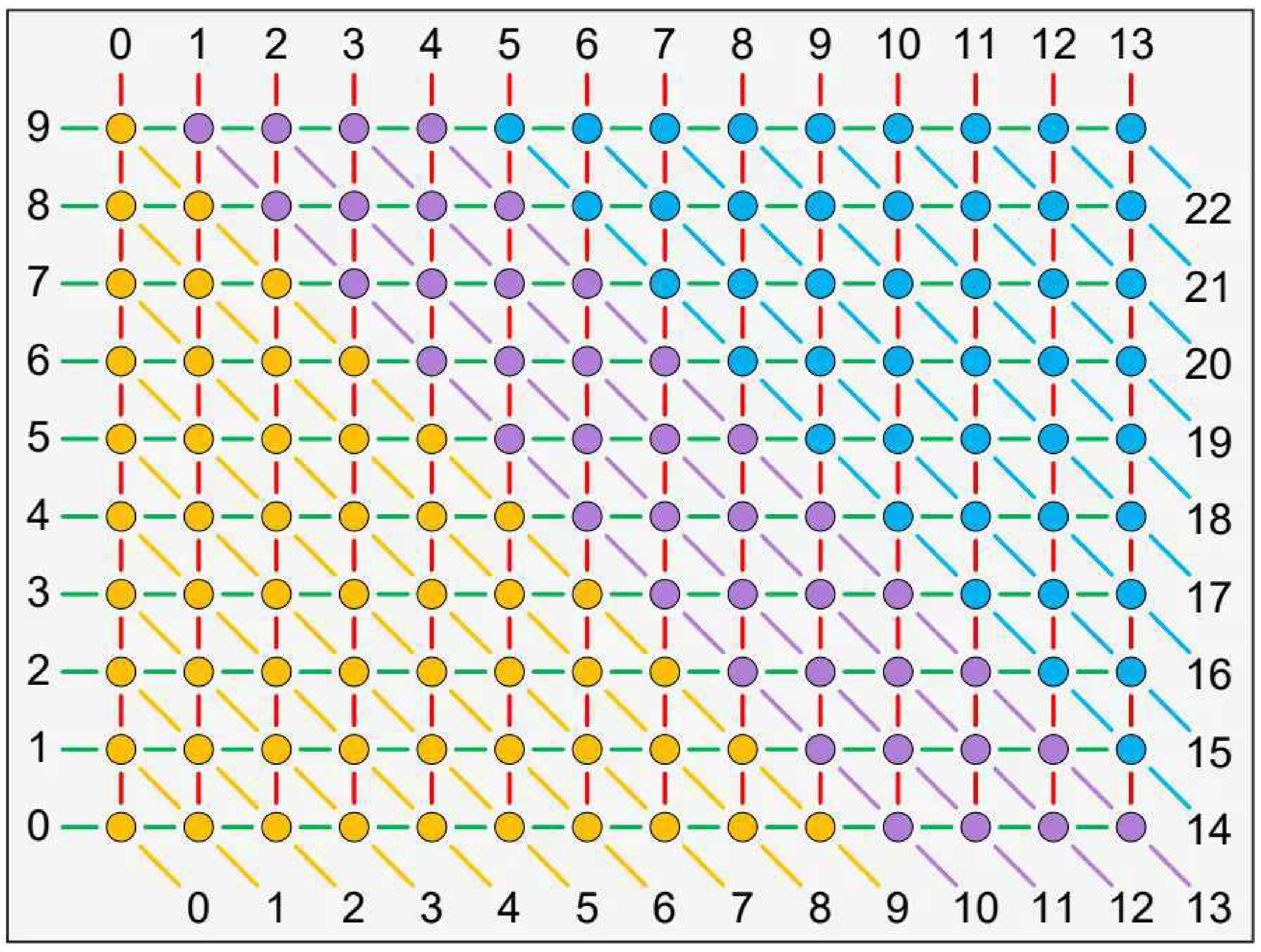
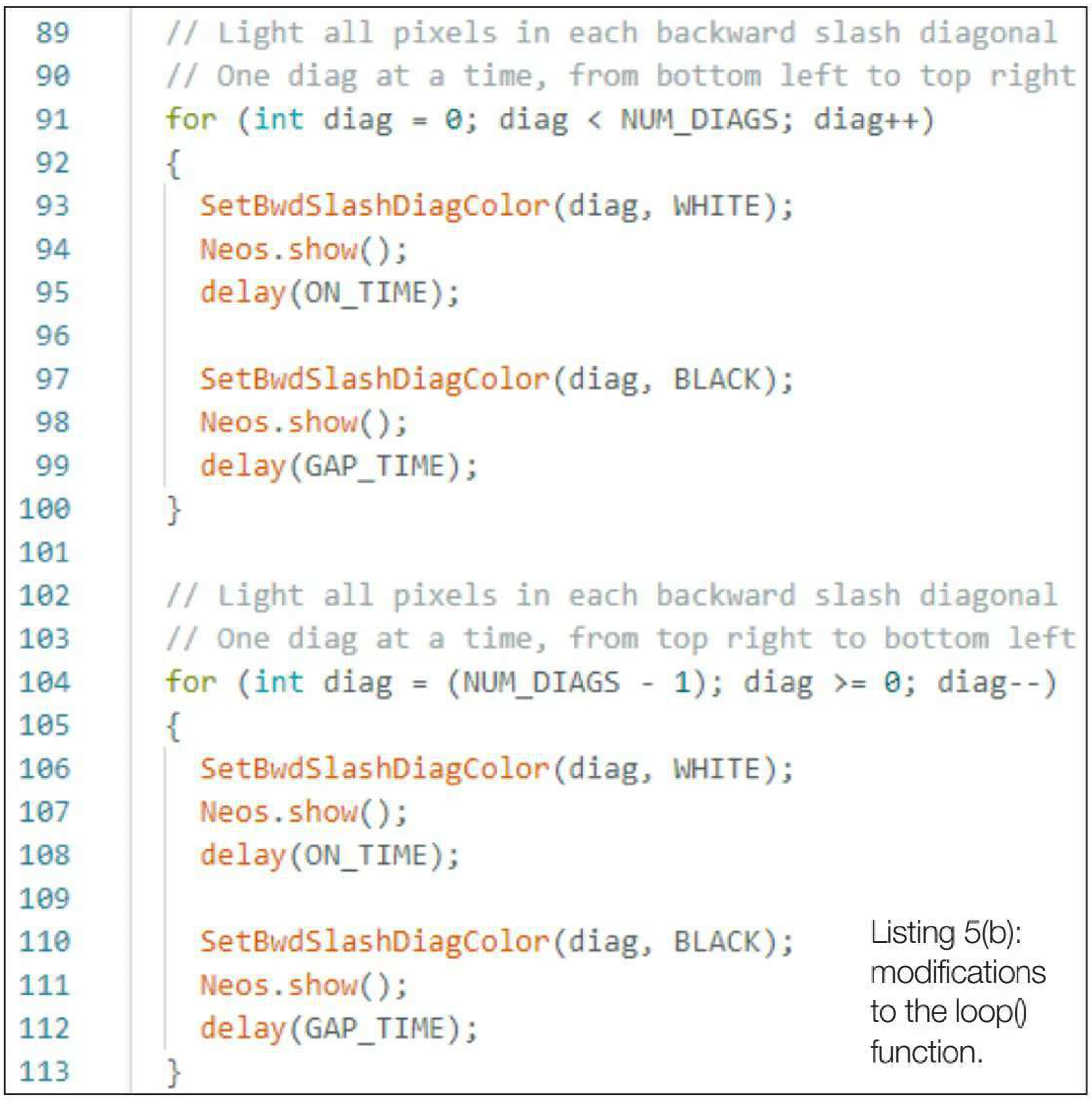


Fig.4: full-size array reflecting the info from our table.



```
 89    // Light all pixels in each backward slash diagonal
 90    // One diag at a time, from bottom left to top right
 91    for (int diag = 0; diag < NUM_DIAGS; diag++)
 92    {
 93      SetBwdSlashDiagColor(diag, WHITE);
 94      Neos.show();
 95      delay(ON_TIME);
 96
 97      SetBwdSlashDiagColor(diag, BLACK);
 98      Neos.show();
 99      delay(GAP_TIME);
100    }
101
102    // Light all pixels in each backward slash diagonal
103    // One diag at a time, from top right to bottom left
104    for (int diag = (NUM_DIAGS - 1); diag >= 0; diag--)
105    {
106      SetBwdSlashDiagColor(diag, WHITE);
107      Neos.show();
108      delay(ON_TIME);
109
110      SetBwdSlashDiagColor(diag, BLACK);
111      Neos.show();
112      delay(GAP_TIME);
113    }
```

Listing 5(b): modifications to the loop() function.

Photo 1: a 7805 linear voltage regulator in the TO-220 package.

Last, but certainly not least, we use the **for()** loop on line 169 to cycle through all the pixels forming our diagonal. This is a prime example of such a loop employing multiple elements in its initialisation and modification sections.

All that remains is to augment our **loop()** function to cycle through the diagonals on the display. This is illustrated in Listing 5(b). In lines 91 to 100, we illuminate the diagonals, one at a time, from bottom-left to top-right. Next, in lines 104 to 113, we reverse the process, illuminating the diagonals, one at a time, from top-right to bottom-left.

I'm running this on my array as we speak - it looks awesome! I can't wait for you to see it running you your own array.

Scotty, I need more power!

I don't know about you, but I'm tired of not being able to light all the pixels on my array simultaneously. That's about to change. To do so, however, we are going to need more power.

Of course, as is usually the case in engineering, there are multiple options open to us. As I've mentioned in previous columns, my chum Joe Farr has already built a prototype version of our retro games console. Before we turn our attention to my own solution, it will be instructive for us to consider the decisions Joe made.

The first key point is that Joe is using a PIC microprocessor from Microchip Technology. A related point is that Joe's PIC runs at 5V, which is the same as his main LED array. This means that Joe's console requires only a single 5V supply. Another consideration is that Joe decided his console should be capable of being powered either from an external source or from internal batteries.

Since the battery-based solution is a limiting factor, that's where Joe started. He opted to use six high-capacity, rechargeable, AA nickel metal hydride (NiMH) batteries, whose combined 6 × 1.2V gave him 7.2V to play with.

Furthermore, Joe decided to employ a linear voltage regulator in the form of a 7805 (Photo 1). Why this particular device? Joe happens to have a drawer full of these little scamps in his workshop. This provides a prime example of how our decisions are sometimes made for us.

A linear voltage regulator is an electronic device that maintains a constant output voltage regardless of variations in the input voltage or load current (within certain limits). It achieves this by dissipating any excess power as heat.

One important term associated with linear voltage regulators is 'dropout voltage'. This is the minimum difference between the input voltage and the output voltage that the regulator requires to maintain a stable, regulated output. The 7805 supplies an output voltage of 5V, and its dropout voltage is 2V, which means its input voltage must be at least 5V + 2V = 7V.

The reason for the 'dropout' nomenclature is that, if the input voltage falls below the required level, the regulator cannot maintain the desired output voltage. Instead, the output will begin to 'drop out' and follow the input voltage, resulting in an unregulated output.

According to its data sheet, the absolute maximum input voltage for the 7805 is 35V, but the recommended input voltage range is 7–25V.

The amount of heat dissipated by a linear voltage regulator is a function of the difference between input and output voltages and how much current we are drawing. If we are pulling a lot of power, these devices can get hot. Hotter than one can touch without shouting "Ouch!". As a result, it's common to attach them to a heatsink (Joe used a strip of metal mounted inside his console).

As seen in Photo 1, the 7805 is a 3-terminal device. Many (but not all) of these are prefixed with "LM", which stands for "Linear Monolithic". These originated with National Semiconductor, which was acquired by Texas Instruments (TI) in 2011. Other versions of this regulator have related names, like the L7805CV shown in the photo.

All the 7805 needs is a couple of extra components in the form of capacitors, and you are 'off to the races'. A typical 7805-based circuit is shown in Fig.5. The voltage rating of the capacitors should be at least a few volts higher than the maximum they are

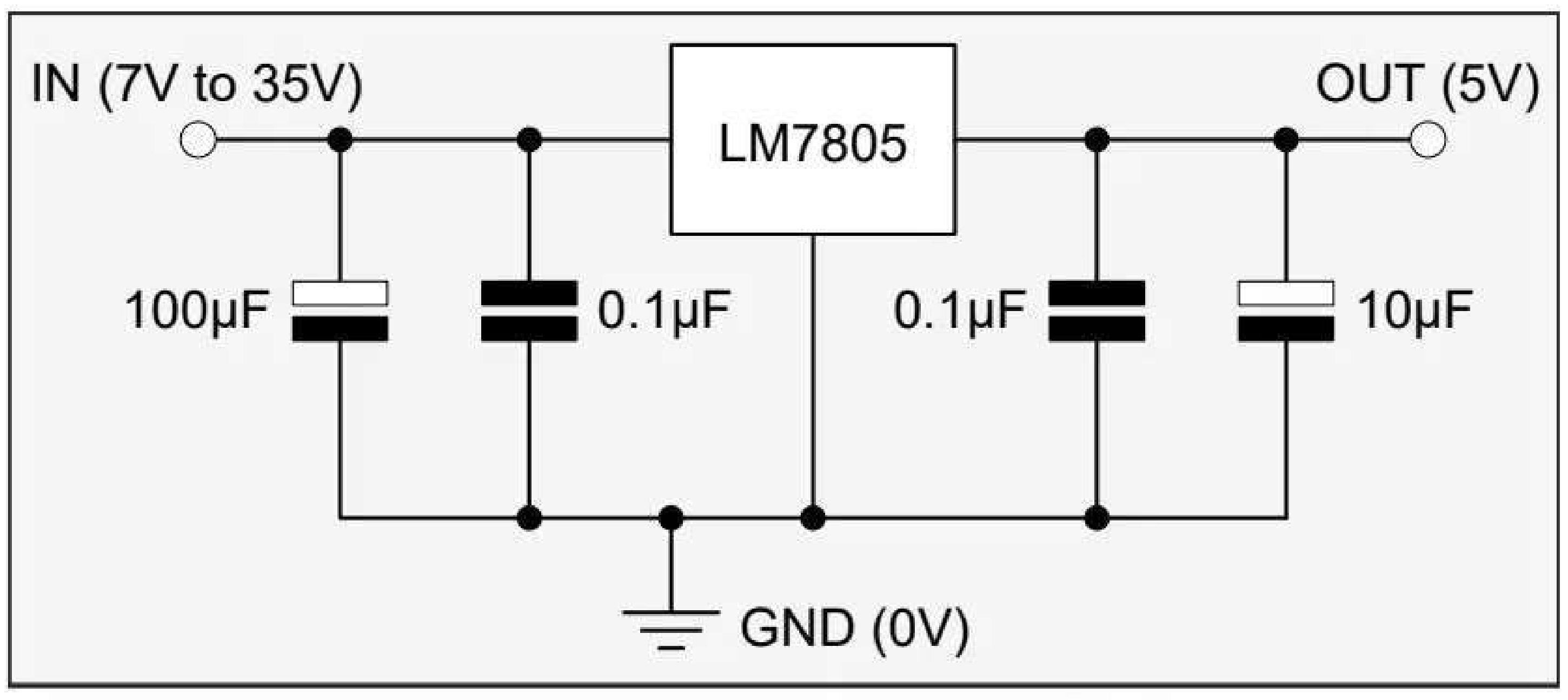


Fig.5: a typical LM7805 regulator-based circuit.

expected to see. As always, consult the data sheet for whatever regulator (and manufacturer) you are using.

A simplified representation of Joe's setup is illustrated in Fig.6. As we previously discussed, the input voltage must be greater than 7V, and this input can either come from an external source or from batteries mounted inside the console.

The 5V output from Joe's regulator feeds both his microcontroller and his LED array. This scheme has the virtues of low cost and simplicity. The biggest downside is the fact that the 7805 can generate a substantial amount of heat when its output is loaded, which will occur if we light all the LEDs on our array, for example.

Back of an envelope

The phrase "back-of-an-envelope calculation" refers to the practice of performing quick, rough calculations on any piece of paper that's close to hand, such as the back of a postage envelope.

In his version of these calculations, Joe allocated himself a total power budget of 1A. Starting with the microcontroller, Joe used an old PIC board from a previous project. This board already had one LED on it. Joe thinks the LED probably consumes more power than the PIC, but he allocated 50mA to the board as a worst-case scenario.

Joe's console boasts a total of six 7-segment displays. As we previously discussed in the February issue, Joe and I are both using dual-digit 7-segment display modules that include 74HC595 shift registers. Also, we're both using three of these modules in our consoles, which means a total of six 7-segment displays and six shift registers.

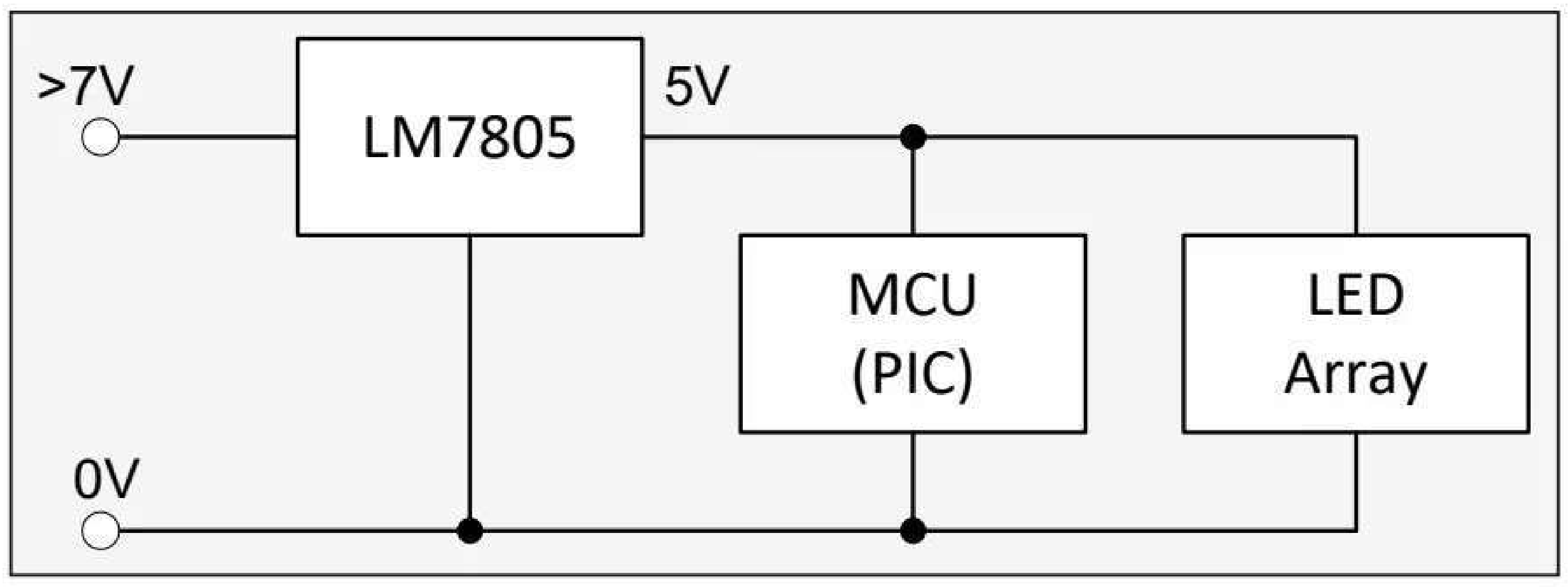


Fig.6: Joe's power supply setup using a linear regulator.

We are assuming that the current used by the shift registers themselves will be negligible for this application. It's hard to get data sheets for these modules, but we are assuming 5mA per segment, which equates to 40mA if all the segments (including the decimal point) are lit on a display. This gives us 80mA per module, or 240mA for all three modules.

Let's add this 240mA to the 50mA for the PIC board and round things up to 300mA, thereby leaving Joe with 700mA for his main array.

Remember that we have 140 pixels, each containing R, G, and B LEDs, each of which is specified as consuming 20mA at full power (so 60mA per pixel). This means that if we actually ran our LEDs at full power, this would equate to 140 × 60mA = 8400mA (8.4A). Wow!

After some experimentation, Joe determined that a value of 16 (0x10) out of a maximum of 255 (0xFF) results in a display that's more than bright enough for gaming without being dazzling. Assuming a linear universe, this means he's running his LEDs at 16 ÷ 255 (or 0x10 ÷ 0xFF) ≃ 6.3%.

This means that if Joe instructs all the pixels on his array to display white light, this will result in 8,400mA × 0.063 ≃ 530mA, well within his 700mA budget.

No restraint!

I must admit that I'm impressed with Joe's self-restraint. Sad to relate, however, I never ran an LED at only 6.3% of its rated value in my life. Whilst I reserve the right to change my mind later, in our existing programs, I'm currently running my LEDs with values of 32 (0x20), which is ~12.5% of their maximum value (that's twice the brightness being used by Joe).

This means that if I'm driving all the pixels on my array white, this will result in a current draw of 8400mA × 0.125 = 1050mA (1.05A). If we add 250mA to cover my six 7-segment displays, this comes to 1300mA (1.3A). I'm going to round this up to 1500mA (1.5A) 'just to be sure'.

Note that I've excluded the Arduino from these calculations because—as we will see—it's covered by a different power budget.

Powering an Arduino

Unlike Joe, I'm using an Arduino Uno R3 to drive my retro games console. Also, unlike Joe, I'm not planning on powering my prototype console using batteries; I'm more than happy to use an external power supply.

There are several ways to power an Arduino Uno R3 (Fig.7). One is to feed 7–12V into the power jack. Also note the bank of header pins marked POWER. For all intents and purposes, the V_{IN} pin on this bank is connected to the positive terminal of the power jack. This means that, as opposed to using the power jack, we can feed 7–12V directly into the V_{IN} pin.

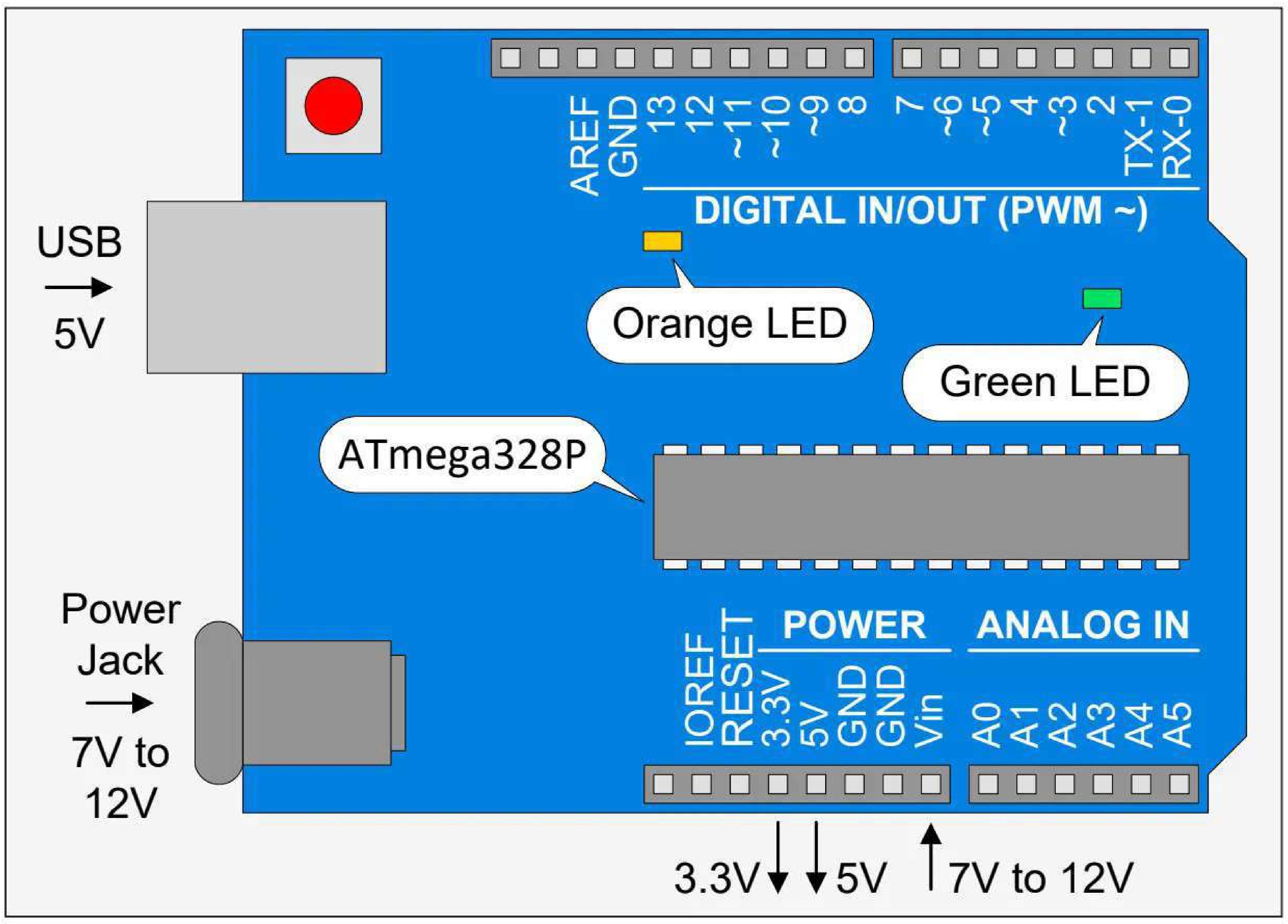


Fig.7: this shows a few different ways you can powering the Arduino.

The Arduino has two onboard regulators. The first takes the voltage from the power jack (or V_{IN} pin) and delivers 5V to the board's internal 5V power rail. In addition to powering the Arduino itself, this 5V power rail is also presented as an output on the POWER section of the Arduino's header pins.

Via its header pin, this 5V supply can be used to drive external circuits, like the breadboard-based clock we created in our *Arduino Bootcamp* columns.

Although the Arduino Uno R3 runs on 5V, many sensors, displays, and other chips use 3.3V. Thus, the 5V power rail feeds the input to a second onboard regulator, which produces 3.3V. This 3.3V supply is also presented as an output on the POWER section of the Arduino's header pins.

Last, but certainly not least, we can power the Arduino via its USB port.

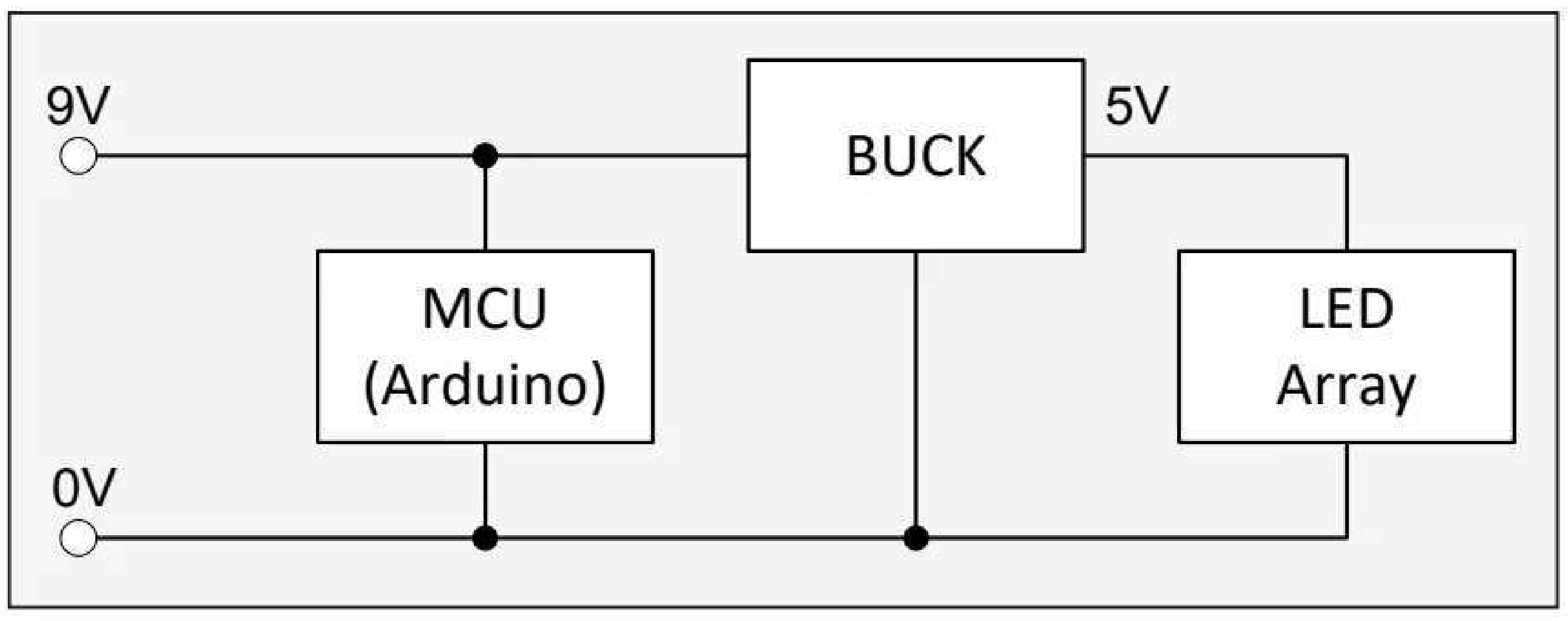


Fig.8: my power supply setup.

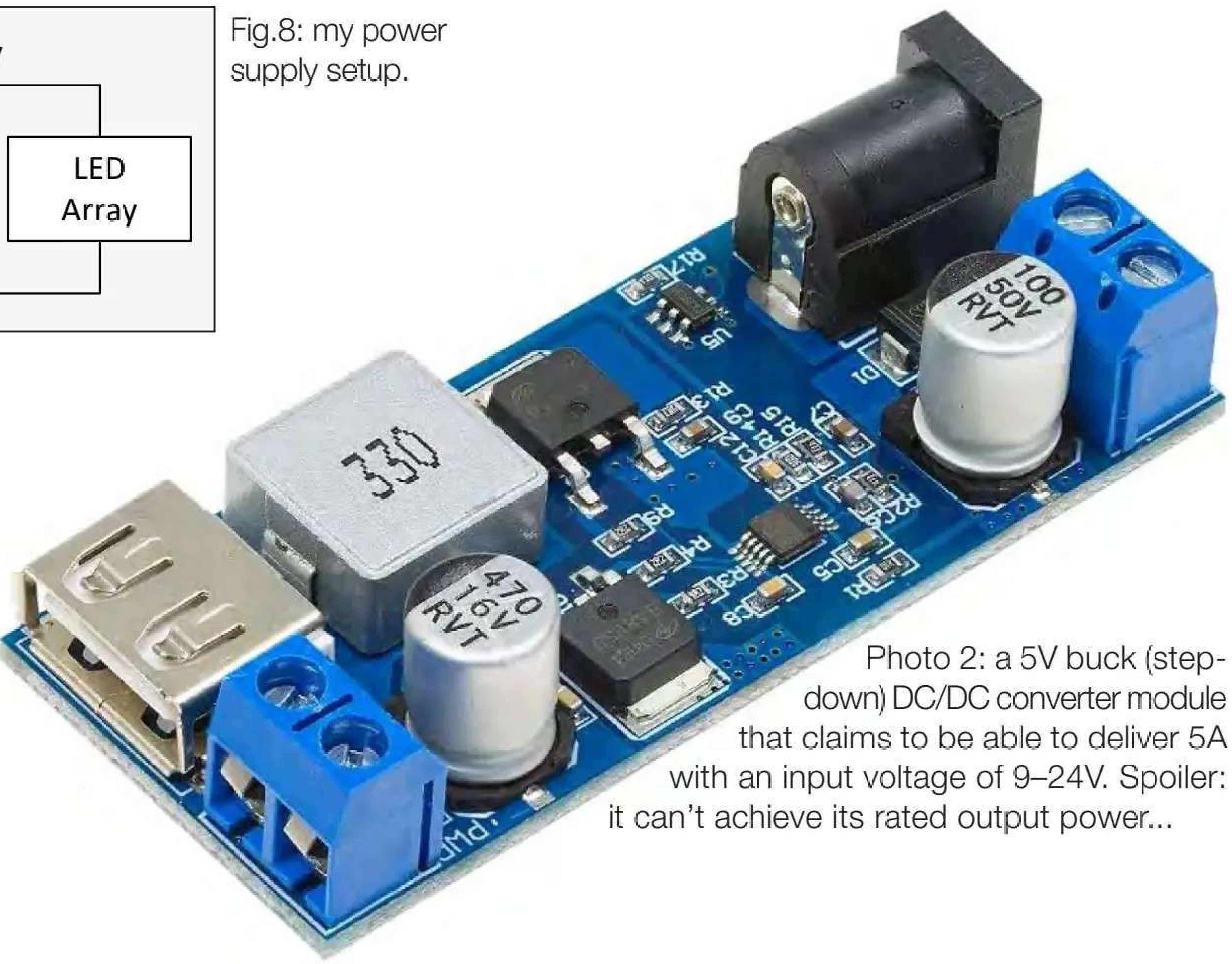


Photo 2: a 5V buck (step-down) DC/DC converter module that claims to be able to deliver 5A with an input voltage of 9–24V. Spoiler: it can't achieve its rated output power...

When USB is connected, the Arduino automatically selects USB as its power source, overriding any external power sources.

Since USB provides a stable 5V supply, this power source bypasses the board's 5V regulator and directly drives the board's 5V power rail. As we already discussed, this powers the Arduino, drives the input to the 3.3V regulator, and appears on the POWER header.

Now, here's the tricky part. Since the 5V pin on the POWER header is directly connected to the board's 5V power rail, as opposed to using it as an output, it is possible to treat it as an input. However, I wouldn't recommend doing this.

The 5V pin bypasses the Arduino's onboard voltage regulators and protection circuits. This means that if the power supply you feed into this pin is unregulated or noisy, it can damage the microcontroller and other components on the board.

Also, if you connect a 5V power source to the 5V pin and simultaneously connect the Arduino to a computer via USB, the two power sources can conflict, possibly damaging your Arduino and/or your computer.

A well-regulated microcontroller

Based on what we just discussed, I decided to feed my console with a 9V power supply. Inside the console, I'm going to feed this 9V signal directly to the Arduino's power jack while prototyping (and to its **V_{IN}** pin once we've started implementing our games cartridge boards).

I'm also going to feed this 9V signal through a 5V regulator and use the output from the regulator to power my LED array and other console components requiring a 5V supply, like my 7-segment displays, for example.

The next question is how to convert my 9V source into the 5V I need to power my array. I could use a linear regulator like Joe, but I'm not keen on generating a lot of heat inside my console.

There are two types of DC-to-DC converters that rely on high-frequency switching to transfer energy efficiently from input to output (there are more, but let's stick with the two most common types). DC stands for 'direct current', which means a current that is steady and flows in one direction only.

A 'boost converter' steps up the input voltage to a higher-level output voltage. By comparison, a 'buck converter' steps down the input voltage to a lower-level output voltage. The main consideration for me is that these devices generate negligible heat (at least, as compared to their linear regulator counterparts).

Why "buck"? I think this is like the way a horse bucks to get someone off their back; in this case, the buck regulator is relieving the circuit of some unnecessary voltage, rather than an unwanted rider.

I found some buck converter modules on Amazon here in the USA (Photo 2). At two for only $15, this seemed like a reasonably good deal. There are equivalents to these modules on Amazon UK (**https://pemag.au/link/ac2m**).

These really are rather handy. On the input side, we have a 2.1mm jack socket (centre pin positive) and a two-way screw terminal block that's wired in parallel with the jack socket. On the output side, we have another a two-way screw terminal block wired in parallel with a USB-A connector, which means you can also use this to charge your USB devices if you wish.

These converters purport to accept anywhere from 9–24V DC as input and generate up to 5A at 5V at the output. Since I require only 1.5A (as per our earlier discussions), these are more than adequate for my design (Fig.8).

All the 0V rails on the Arduino, buck converter, and LED array must be connected (0V is equivalent to GND for the purposes of these discussions). The reason why these must all be connected was discussed in detail in the September 2024 issue.

One point that may prove puzzling to a beginner involves the buck converter. Let's look at the bottom of this board as shown in Photo 3 (remember this is my board, yours may vary slightly).

The **V_{IN}+** and **V_{IN}-** annotations correspond to the screw terminal block on the input side of the board. The 5V and **GND** annotations correspond to the screw terminal block on the output side of the board. The point is that the **V_{IN}-** and **GND** pins are directly connected on the board, and these correspond to the 0V rail in Fig.8.

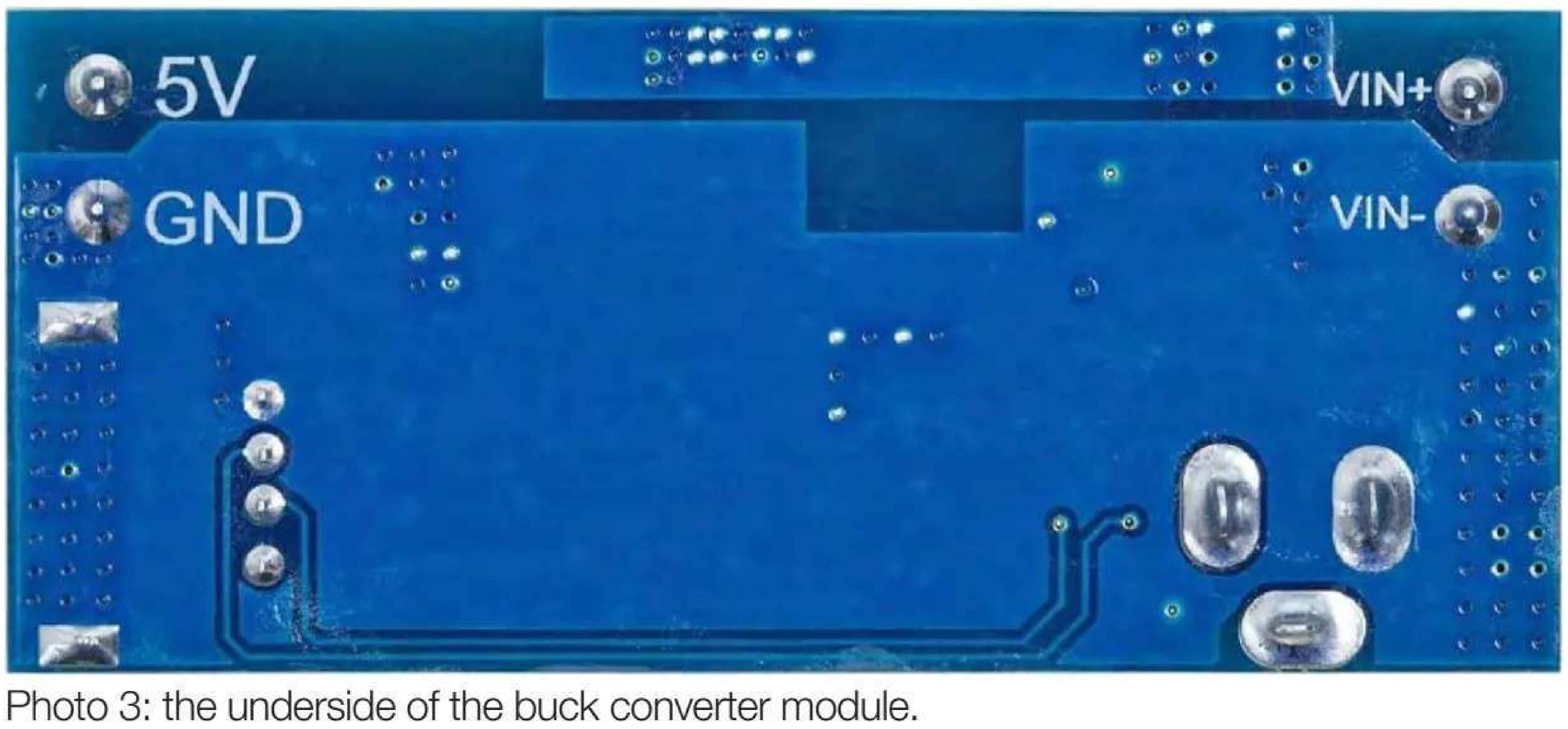


Photo 3: the underside of the buck converter module.

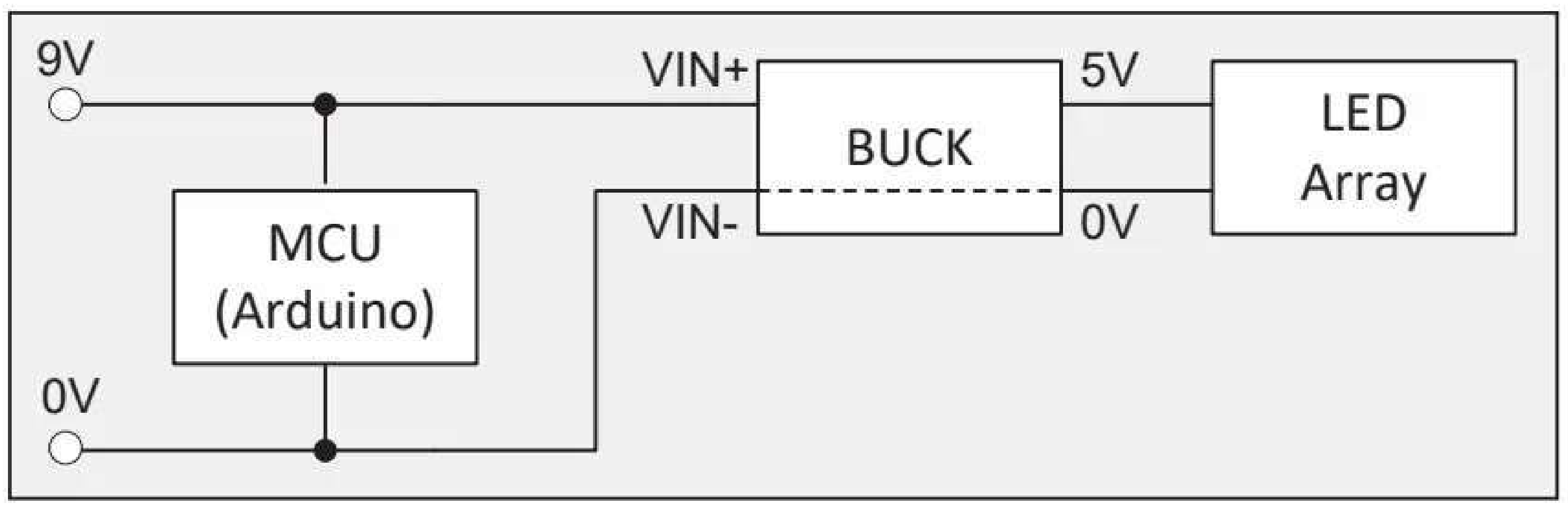


Fig.9: another way of looking at my power supply arrangment.

What this means is that we could redraw my setup as illustrated in Fig.9. Although they look a little different, the circuits in Fig.8 and Fig.9 are functionally the same.

Too good to be true

"Caveat emptor" is a Latin phrase that translates to "let the buyer beware". It refers to the fact that people can and will try to sell you any old nonsense; it's up to you to make sure you're getting what you want.

I must admit that when I saw the aforementioned buck converter modules—especially their claim to be able to supply up to 5A at 5V—I thought, "that sounds too good to be true".

Joe agreed, so he purchased some similar modules of his own for testing purposes. Joe has a pretty handy piece of test equipment called a 'DC electronic load'. This is designed to 'sink' current and absorb power from a power source. It's used to simulate a device that would draw electricity from that source, thereby allowing the user to test the performance of power supplies, batteries, solar panels, and other energy sources under different load conditions.

When Joe applied his electronic load to the output of his DC-to-DC converter, he found it held steady at 5V up to a load of about 1.7A. However, the output sagged to 4.7V when the load reached around 1.8A, and it started to fall off rapidly (ie, collapsed) by the time the load reached approximately 2.1A.

On the one hand, this is disappointing compared to its claims of delivering 5A (you can't trust anyone these days). On the other hand, since I've set my maximum power budget at 1.5A, my converter will satisfy my requirements, even if it has the same limitations as Joe's modules.

Note that the output current likely depends on the input voltage, but if it can't even deliver 2A with a 12V input, there's no way it will manage 5A even with the maximum 24V supply voltage applied.

As one final point, Joe's electronic load is an expensive piece of professional equipment, but cheaper hobbyist versions are available. I just found one on Amazon here in the USA for only $42.99 (https://pemag.au/link/ac45), which is around £34.50. This isn't something most of us need on a daily basis, but I can see how it could come in handy one day, so I've added one to my birthday wish list.

I've got the power!

Returning to my setup (Fig.9), the next thing I'm going to need is a 9V DC power supply that can deliver at least 2A. I know we set the maximum at 1.5A earlier, but I always like to give myself some headroom.

I could purchase yet another standalone power supply, like the 9V 2A unit I just found on Amazon in the UK for about £10 (https://pemag.au/link/ac46). If you decide to go this route, make sure to buy the centre-positive version, which means the centre pin of the barrel jack carries the positive voltage while the outer sleeve has the negative voltage.

This is because the power jacks on both the Arduino Uno and our DC-to-DC converter module are centre-positive. Connecting a centre-negative supply will either do nothing or do something very bad, depending on how well designed the powered device is.

To be honest, I've bought loads of different power supplies over the years, but I still never seem to have the one I need. Joe has a couple of bench power supplies in his workshop, but I don't have the room for one of these in my study at home (which is where I'm writing these columns).

I don't know why it took me so long, but I looked on Amazon here in the USA and found an awesome bench power supply for only about $45 (https://pemag.au/link/ac47 & Photo 4). I just found the same unit on Amazon UK for £50 (https://pemag.au/link/ac48).

This can supply up to 30V and up to 10A (you set the voltage you want and the maximum current you wish the supply to… well, supply). When in operation, it reports the actual output voltage along with how much current your load is sinking, and even the amount of power it's delivering.

This little rascal comes with a bunch of bells and whistles. Best of all, it's only around 3.5 inches (9cm) wide, 6 inches (15cm) tall, and 7.5 inches (19cm) deep, which means it occupies a negligible amount of space on my desk.

NANKADF
Programmable DC Power Supply
C.V
30.00 V
C.C
10.00 A
OCP
OUT
300.0 W
USB
Output
Quick Charge
VOLTAGE
POWER
Push/Hold OCP NO/OFF
CURRENT
+

Photo.4: I love my new bench power supply.

One step at a time

I always like to take things step by step. I'm going to describe what I did with my setup; you can vary these steps to reflect your own situation as required.

First, with its output disabled, I set my bench power supply to 9V DC and limited the output current to 250mA (just in case). Then I connected two pairs of black (0V) and red (5V) leads to the black and red terminals on my power supply. The other ends of these leads were connected to 2.1mm power jack plugs.

Before I connected these plugs to anything, I activated the power supply's output and used my multimeter to confirm I was seeing 9V on each jack's inner connector.

Then I disabled the power supply's output, plugged one of my cables into my DC-to-DC converter's input jack, and re-enabled the power supply's output. The power supply showed it was delivering 9V and 5mA, so now we know that the converter draws 5mA even when it's not driving anything. I used my multimeter to confirm that I was seeing 5V at the output of the converter.

I want to protect my array at all costs. So, before I did anything else, I disconnected it from my Arduino.

Next, I connected the USB cable from my PC to the Arduino, and loaded the standard "Blink" program that comes with the Arduino (you can use the Arduino IDE to find this under the File → Examples → 0.1.Basics → Blink menu item). This repeatedly flashes the Arduino's onboard LED on for a second and off again for a second.

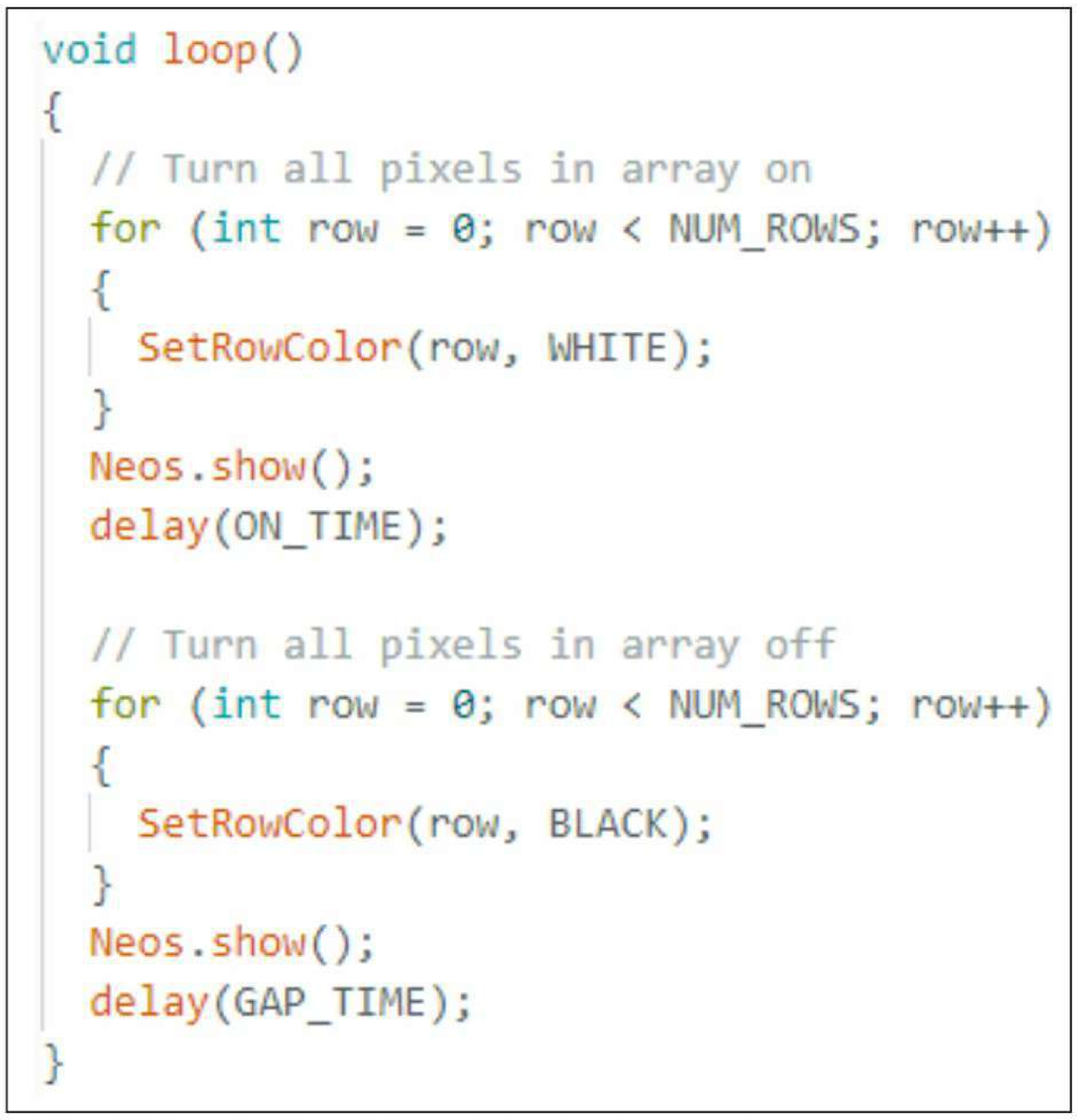
Once the Blink program was running, I unplugged the USB cable from the Arduino, disabled the power supply's output and plugged the second cable from the power supply into the Arduino's input jack. When I reactivated the power supply's output, the Arduino powered up and returned to running its Blink program. Also, the power supply now showed it was delivering 9V and 55mA.

Subtracting the 5mA for the converter shows the Arduino is pulling 50mA, which is as expected.

I disabled the power supply's output, reconnected the LED array to the Arduino and enabled the power supply's output once again. Remember that I'm still running the Blink program. Also, all the pixels in the array power up in their off condition.

Now, my power supply shows it's delivering 9V and 125mA. This is interesting. If we subtract 5mA for the DC-to-DC converter (whose output we still aren't using) and 50mA for the Arduino, this leaves 70mA being consumed by the array. Since we have 140 pixels, this means each pixel consumes half a milliamp, even when it's not displaying anything. I must admit, I wasn't expecting that.

Next, I plugged the USB cable back into the Arduino. As we discussed earlier, when a USB cable is connected, the Arduino automatically selects USB as its power source, overriding any external power sources. Thus, it wasn't surprising that the power supply returned to delivering only 5mA to power the converter.



```
void loop()
{
  // Turn all pixels in array on
  for (int row = 0; row < NUM_ROWS; row++)
  {
    SetRowColor(row, WHITE);
  }
  Neos.show();
  delay(ON_TIME);

  // Turn all pixels in array off
  for (int row = 0; row < NUM_ROWS; row++)
  {
    SetRowColor(row, BLACK);
  }
  Neos.show();
  delay(GAP_TIME);
}
```

Listing 6(a): activating all the pixels simultaneously.

I reloaded our current latest-and-greatest program into the Arduino—the one that displays single rows, columns, and diagonals (in the file named **CB-Apr25 Code-05.txt**). This program continued to work as expected.

Finally, I disabled the output from the power supply. I left the control signal from digital pin 12 on the Arduino feeding the array, but I removed the red (5V) and black (0V) leads from the Arduino—the ones that were powering the array—and connected these to the output from the DC-to-DC converter.

When I re-enabled the output from the power supply, everything carried on working as before. The only difference was that the Arduino was being powered with 9V directly from the power supply, while the array was being powered by the 5V from the DC-to-DC converter.

It's go time!

Now that we have sufficient power, we can activate all the pixels in our array at the same time. I've modified our existing program to do this. The new version is in the file named **CB-Apr25 Code-06.txt**.

First, I changed the definitions for ON_TIME and GAP_TIME to be 2000 and 1000 or two seconds and one second, respectively.

Next, I changed the loop() function

Here I am, using my new power supply for the first time!

to repeatedly turn all the pixels on (for `ON_TIME`) and off again (for `GAP_TIME`). This portion of the program is shown in Listing 6(a).

Remember that we are currently driving our pixels at 12.5% of their full power. The data sheet says full power is 60mA per pixel, so when our 140-pixel array is fully lit at 12.5%, that would equate to 140 × 60mA × 0.125 = 1050mA. If we add 5mA for the converter, 50mA for the Arduino, and 70mA for the intrinsic current draw of the array with no pixels lit, this should give a grand total of 1175mA.

Interestingly enough, my bench power supply reports a maximum value of only 450mA when all the pixels are driven white at 12.5%. That's less than half of the expected current draw. Part of the reason for this is that 450mA @ 9V is 4.05W, and if the buck converter is, say, 90% efficient, that equates to a 5V output of 4.05W ÷ 5V = 810mA.

In other words, you get more current out of the buck converter than you put in because you're reducing the voltage by more than you're reducing the power. Still, it's a little on the low side. I will investigate this further, and I'll report back next month.

For shame!

I fear I must hang my head in shame. In my previous column, I promised that we would be performing some more experiments with our eight-segment bar graph display in conjunction with a shift register. Sadly, I got carried away with our retro games console, but we will certainly do this in my next column (no, my fingers aren't crossed behind my back—why would you ask?).

Until that frabjous day, if you have any thoughts that you'd care to share on anything you've read here, please feel free to drop me an email at max@clivemaxfield.com. Until next time, have a good one! **PE**

Useful Bits and Pieces from Our Arduino Bootcamp series

Arduino Uno R3 microcontroller module	https://pemag.au/link/ac2g
Solderless breadboard	https://amzn.to/3O2L3e8
8-inch (20cm) jumper wires (male-to-male)	https://amzn.to/3O4hnxk
Long-tailed 0.1-inch (2.54mm) pitch header pins	https://pemag.au/link/ac2h
LEDs (assorted colours)	https://amzn.to/3E7VAQE
Resistors (assorted values)	https://amzn.to/3O4RvBt
Ceramic capacitors (assorted values)	https://pemag.au/link/ac2i
16V 100µF electrolytic capacitors	https://pemag.au/link/ac2j
Momentary pushbutton switches	https://amzn.to/3Tk7Q87
Kit of popular SN74LS00 chips	https://pemag.au/link/ac2k
74HC595 8-bit shift registers	https://pemag.au/link/ac1n

Other stuff

Soldering guide	https://pemag.au/link/ac2d
Basic multimeter	https://pemag.au/link/ac2f

Components for Weird & Wonderful Projects, part 1

4-inch (10cm) jumper wires (optional)	https://pemag.au/link/ac2l
8-segment DIP red LED bar graph displays	https://pemag.au/link/ac2c

Components for Weird & Wonderful Projects, part 2

144 tricolour LED strip **(required)**	https://pemag.au/link/ac2s
Pushbuttons (assorted colours) **(required)**	https://pemag.au/link/ac2t
7-Segment display modules **(recommended)**	https://pemag.au/link/ac2u
2.1mm panel-mount barrel socket **(optional)**	https://pemag.au/link/ac2v
2.1mm inline barrel plug **(optional)**	https://pemag.au/link/ac2w
25-way D-sub panel-mount socket **(optional)**	https://pemag.au/link/ac2x
25-way D-sub PCB-mount plug* **(optional)**	https://pemag.au/link/ac2y
15-way D-sub panel-mount socket **(optional)**	https://pemag.au/link/ac2z
15-way D-sub PCB-mount plug^ **(optional)**	https://pemag.au/link/ac30

* one required for each game cartridge you build
^ one required for each auxiliary control panel you build

Components for Weird & Wonderful Projects, part 3

22 AWG multicore wire kit	https://pemag.au/link/ac3m
20 AWG multicore wire kit	https://pemag.au/link/ac3n

Components for Weird & Wonderful Projects, part 4

5V Buck converter module	https://pemag.au/link/ac2m
9V 2A power supply **(optional)**	https://pemag.au/link/ac46
Bench power supply **(optional)**	https://pemag.au/link/ac48

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Using a MEMS Microphone as a Reference Microphone

by Phil Prosser

MEMS (micro-electromechanical system) microphones have advantages over electret mics, such as operating at ultrasonic frequencies. They also have good frequency response characteristics, so you can use them as reference microphones, as described in this article.

We received some Knowles SPU0410LR5H MEMS microphone elements from a kind reader named Richard Stone. They were sent to determine their suitability for use as calibrated microphones.

That was prompted by our Calibrated Measurement Mic project that was published in the September 2024 issue. It used inexpensive electret capsule microphones (ECMs). It used compensation and calibration to provide a flat frequency response, allowing those microphones to be used as measurement devices, eg, to plot the frequency response of a loudspeaker.

The MEMS microphones we received are tiny (3.76 × 2.95mm) and connect to a PCB via under-chip pads. They also require a hole in the PCB that's used as the aperture for the microphone, so they must be soldered to a PCB designed explicitly for them. Soldering them would be tricky for most of our readers. They are surprisingly inexpensive at only around $1 each (less in quantity).

Happily, it turns out that you can buy these microphones already assembled to a board from TeensyBat: https://pemag.au/link/abt5

That is just one example; there are quite a few suppliers of similar 'carrier boards'. The ones we tested came mounted on 7mm circular PCBs.

The Knowles MEMS microphone needs a 1.5-3.6V DC power supply and provides an AC output. As a result, they can be connected to our Calibrated Microphone board but some minor modifications are required. These involve adding a 3.3kΩ series resistor and 3.3V zener across the microphone power supply to obtain a suitable voltage, as shown in the revised circuit diagram, Fig.1.

To do this on the SMD version of the PCB, you have to cut the track between capacitor C6 (10μF) and resistor R4 (100kΩ), which is small but not too fiddly.

This is shown in Fig.2, along with the added 3.3kΩ resistor and microphone wiring. If using an SMD resistor, it can be soldered across the pads spanning the cut location, although adding a miniature through-hole resistor, as shown, is easier.

The equivalent changes for the through-hole version of the PCB are shown in Fig.3. In both cases, the rear of the 7mm round microphone PCB mentioned above is illustrated for the wiring. However, you might prefer to route the wires from the

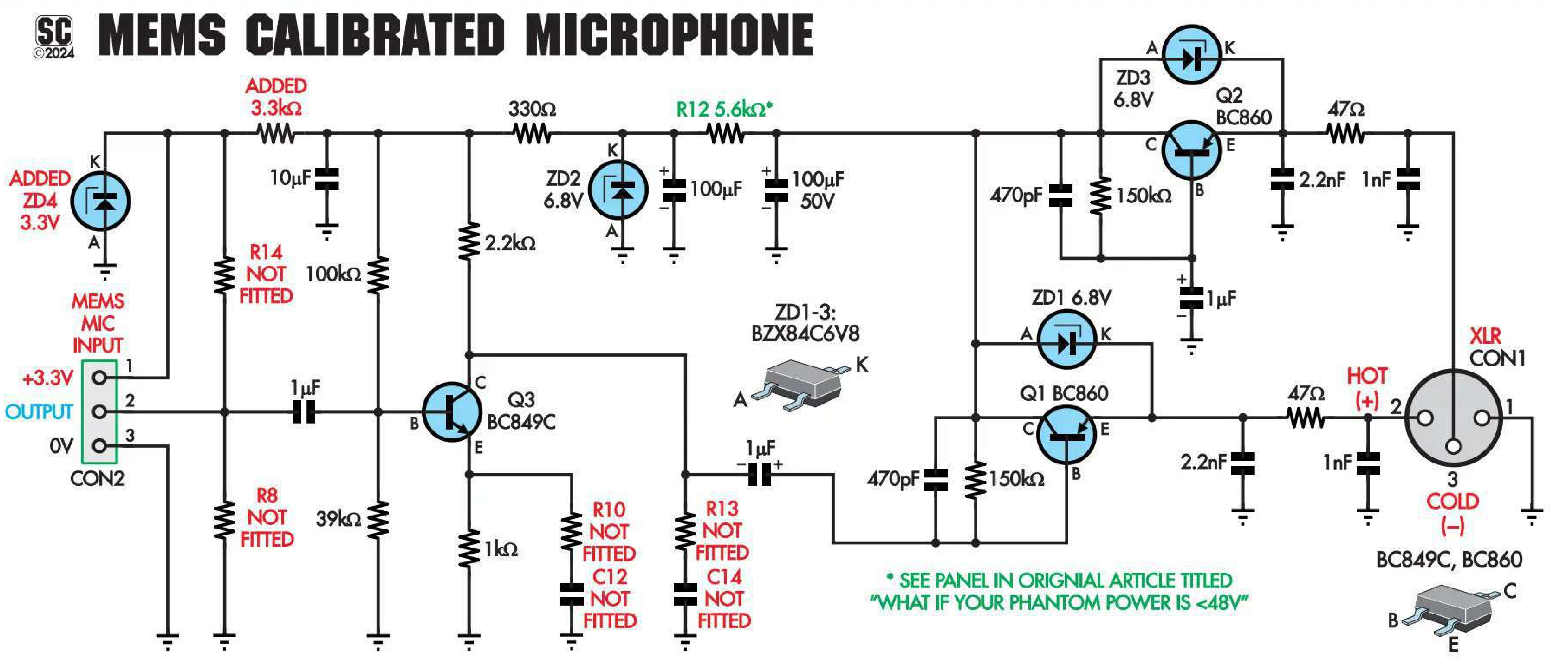
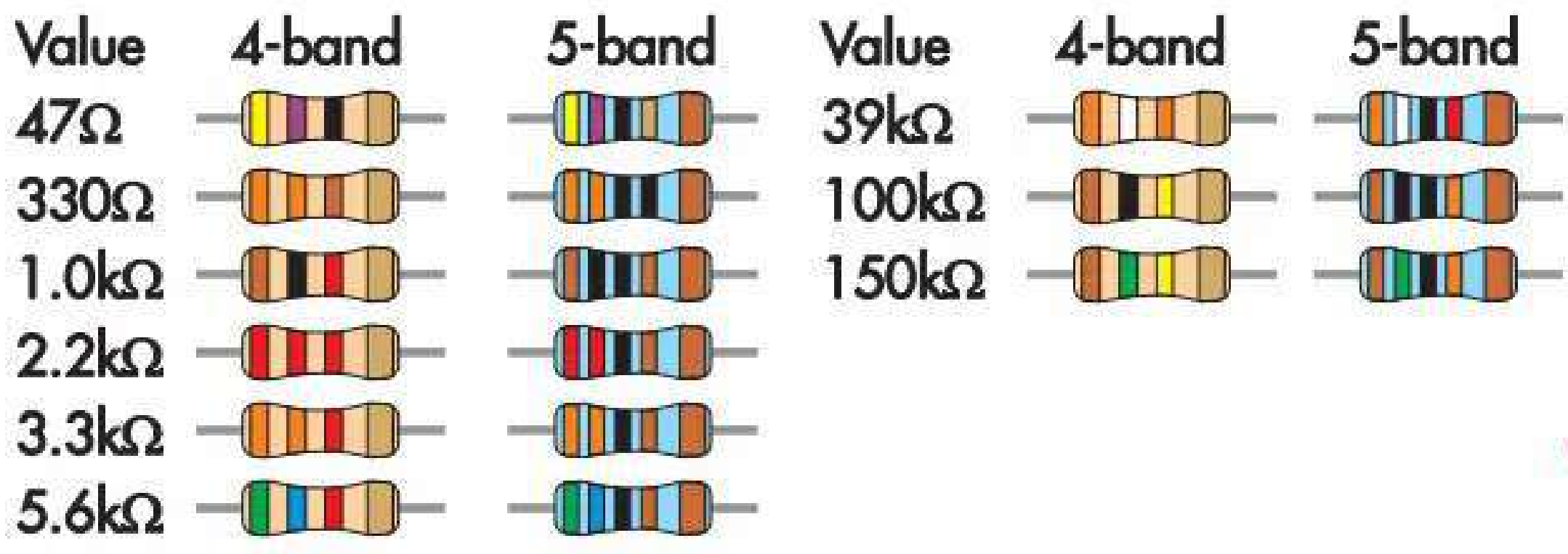

SC ©2024
MEMS CALIBRATED MICROPHONE
ADDED 3.3kΩ
330Ω
R12 5.6kΩ*
ZD3 6.8V
Q2 BC860
47Ω
ADDED ZD4 3.3V
10μF
ZD2 6.8V
100μF
100μF 50V
470pF
150kΩ
2.2nF
1nF
R14 NOT FITTED
100kΩ
2.2kΩ
1μF
MEMS MIC INPUT
+3.3V
OUTPUT
0V
CON2
ZD1-3: BZX84C6V8
ZD1 6.8V
Q1 BC860
XLR CON1
1μF
Q3 BC849C
47Ω
HOT (+)
2.2nF
1nF
470pF
150kΩ
3 COLD (–)
R8 NOT FITTED
39kΩ
1kΩ
R10 NOT FITTED
C12 NOT FITTED
R13 NOT FITTED
C14 NOT FITTED
* SEE PANEL IN ORIGNIAL ARTICLE TITLED "WHAT IF YOUR PHANTOM POWER IS <48V"
BC849C, BC860
A K B C E

Fig.1: the changes required to the original Calibrated Microphone preamp circuit are minimal. R8, R14 and the four compensation components are not fitted, a 3.3kΩ resistor replaces the track between pin 1 of CON2 and the 10μF capacitor, and a 3.3V zener across pins 1 and 3 of CON2 limits the microphone's supply voltage to a safe level.



Value	4-band	5-band
47Ω		
330Ω		
1.0kΩ		
2.2kΩ		
3.3kΩ		
5.6kΩ		
39kΩ		
100kΩ		
150kΩ		

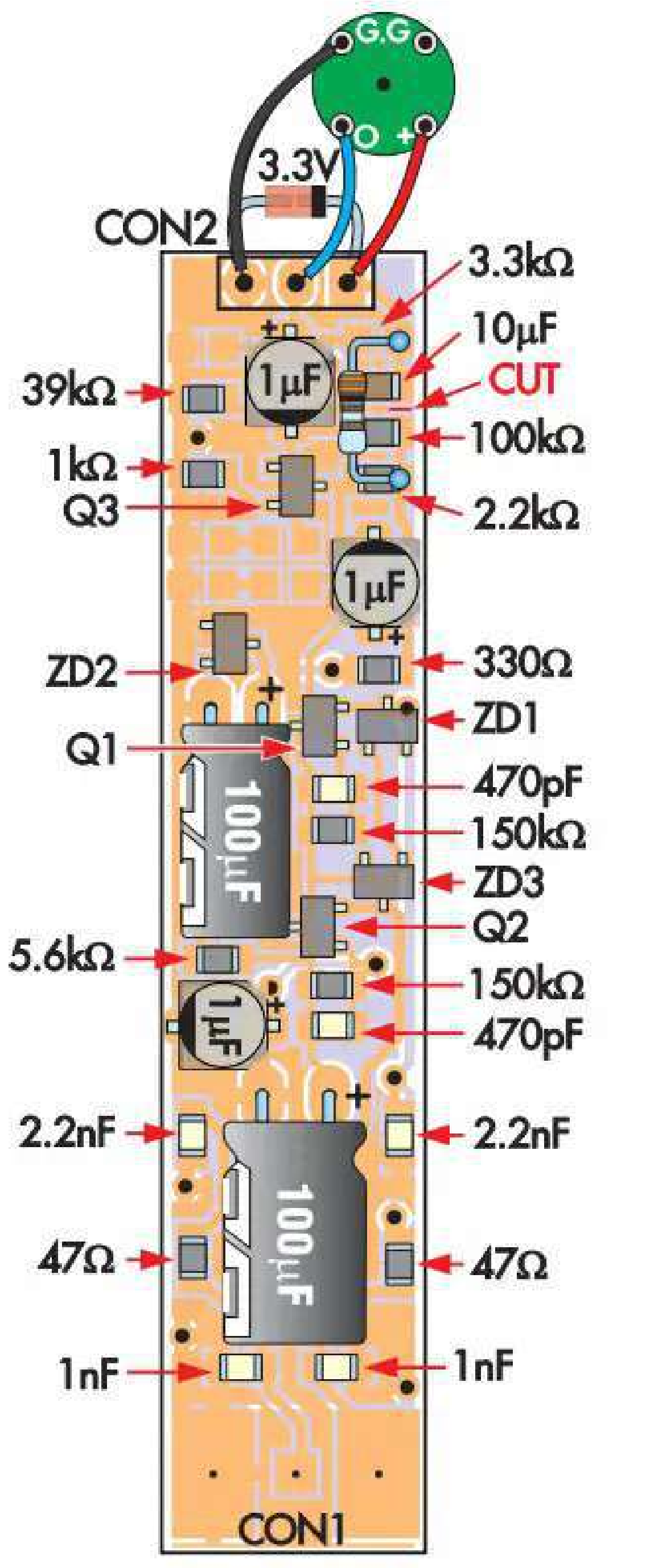

G.G
O
+
3.3V
CON2
3.3kΩ
10µF
CUT
39kΩ
1µF
100kΩ
1kΩ
Q3
2.2kΩ
1µF
330Ω
ZD2
ZD1
Q1
100µF
470pF
150kΩ
ZD3
Q2
5.6kΩ
1µF
150kΩ
470pF
2.2nF
2.2nF
100µF
47Ω
47Ω
1nF
1nF
CON1

Fig.2: this shows how to assemble the SMD version of the PCB and wire it up to the MEMS microphone. The through-hole 3.3kΩ resistor shown could be replaced with an SMD resistor across the cut section of track (soldered on top of the leads of the other components). Your microphone board might differ from the one shown here, so be careful to wire it up correctly.

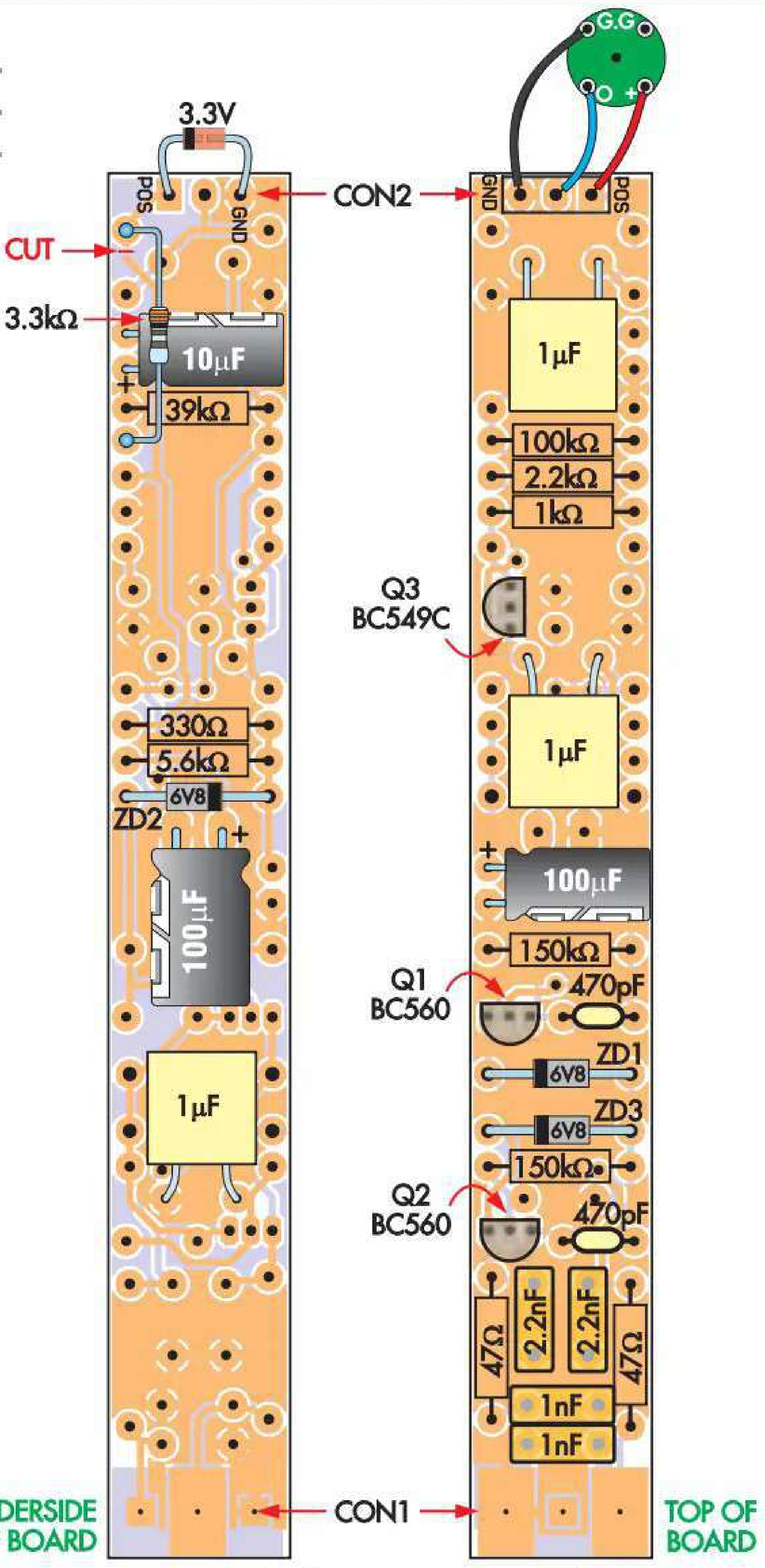

3.3V
POS
GND
CON2
CUT
3.3kΩ
10µF
39kΩ
330Ω
5.6kΩ
6V8
ZD2
100µF
1µF
UNDERSIDE OF BOARD
CON1
G.G
O
+
GND
POS
1µF
100kΩ
2.2kΩ
1kΩ
Q3 BC549C
1µF
100µF
150kΩ
Q1 BC560
470pF
6V8
ZD1
6V8
ZD3
150kΩ
Q2 BC560
470pF
47Ω
2.2nF
2.2nF
47Ω
1nF
1nF
TOP OF BOARD

Fig.3: as with the SMD version, several components are left off the through-hole version of the PCB, one track is cut and a resistor and zener diode are added. Note how the striped end of the extra zener diode goes to the positive (supply) terminal of CON2.

other side to keep the area with the sensing hole clear. The pads labelled "G" are ground, "O" is the output and "+" is the positive supply.

Note that while both of our boards have mounting locations for frequency compensation parts (two resistors and two capacitors), we leave them off for this microphone as it does not require compensation.

The MEMS microphone connected this way works a treat. The resulting 'calibration curve' is shown in Fig.4. The cyan curve is the frequency response of this microphone, while the Dayton EMM-6 reference mic we used for the original project is in red. The calibration data we have for the Dayton unit only runs from 20-20000Hz, so I cut the measurements off there.

Note that the speaker used for this test was rolling off in its response at low frequencies, so the measurements are noisy down low.

The measured response is entirely consistent with published data. The MEMS microphone's output level is much higher than the Dayton microphone, and per the data sheet, the SPL (sound pressure level) limit is not that high, so you will be limited in making near-field measurements or dealing with high SPLs.

In terms of calibration, if you only want to measure up to 10kHz, you can probably ignore the calibration

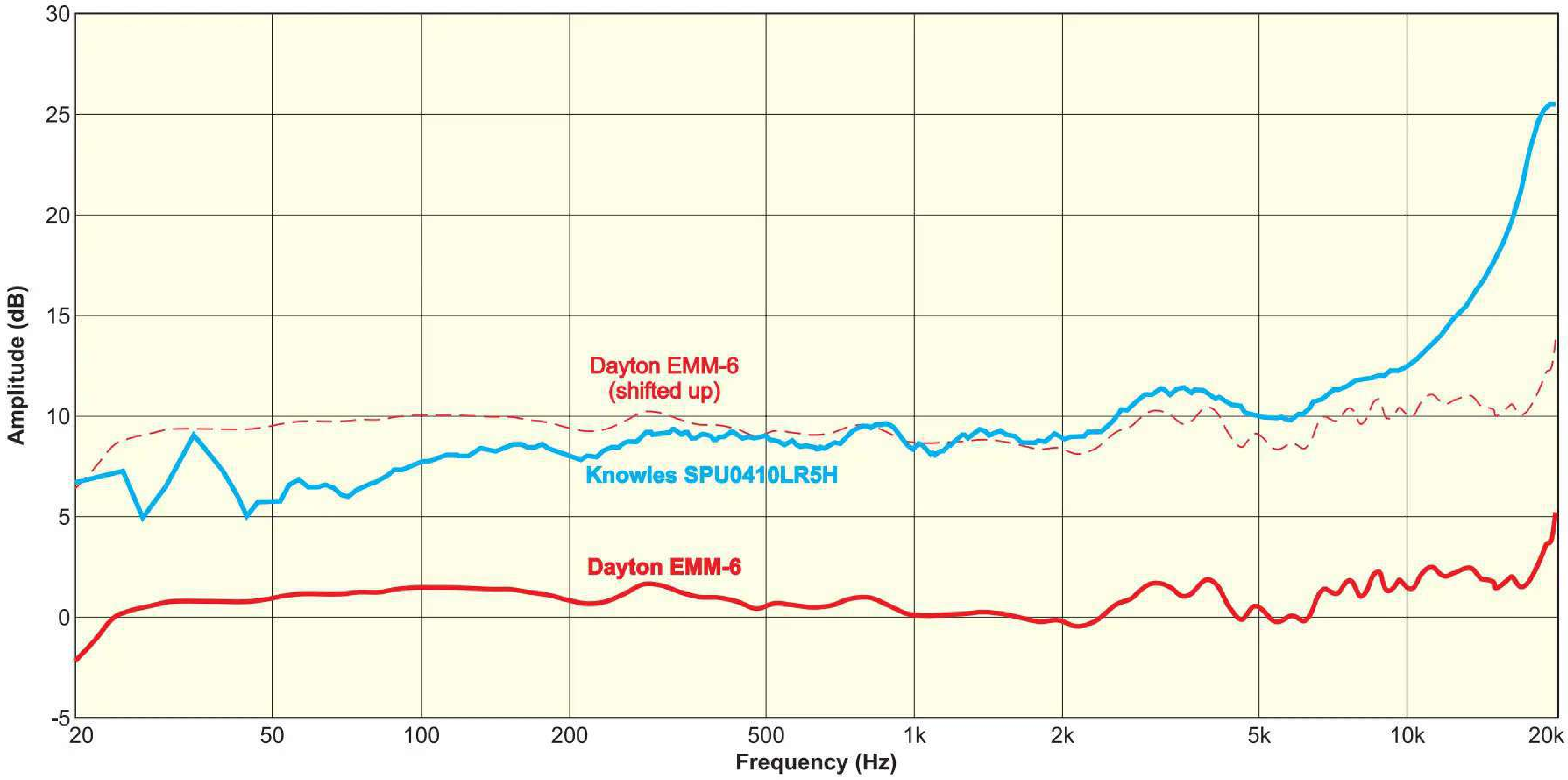
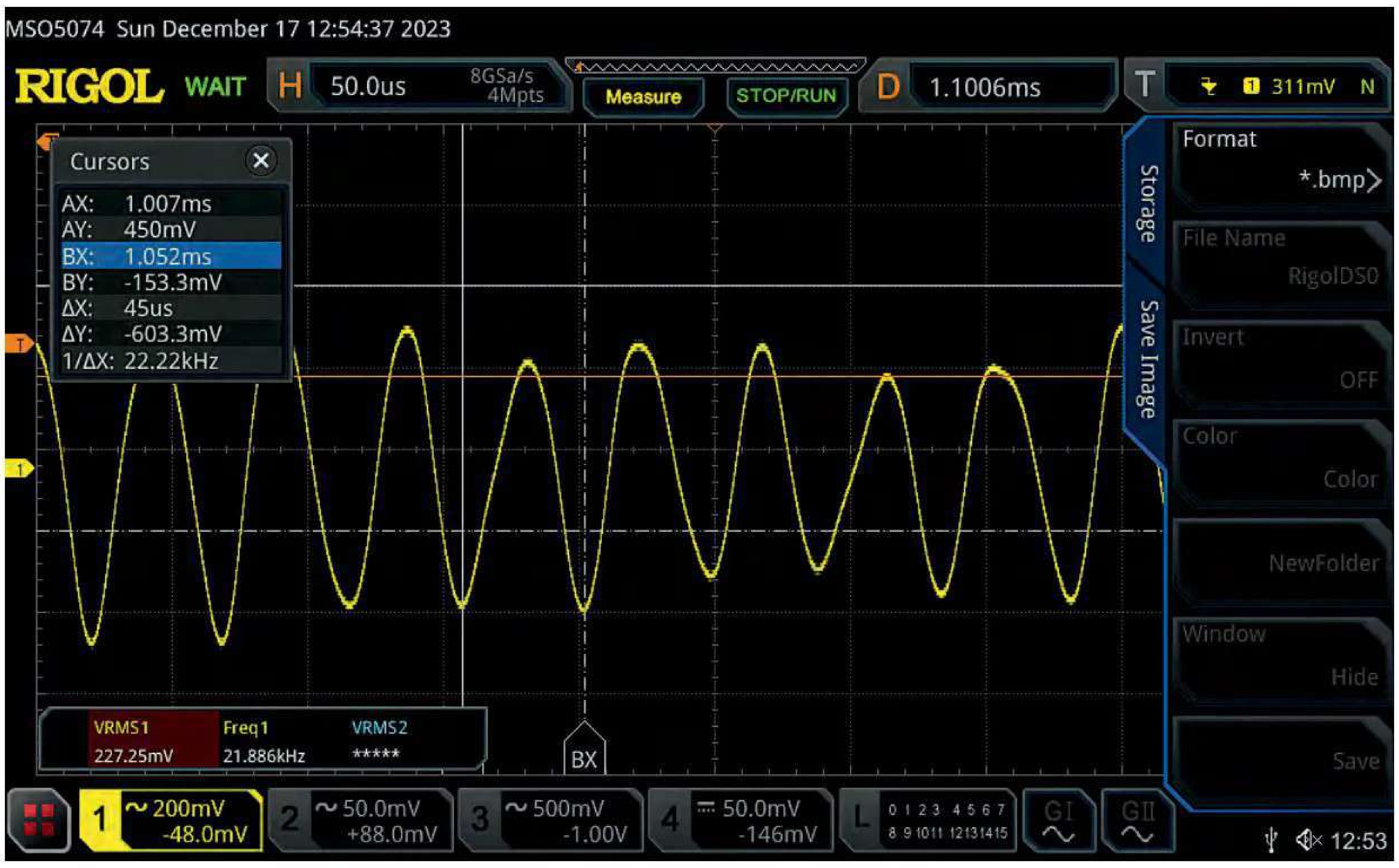


Fig.4: the raw frequency response of the Knowles MEMS microphone (blue) compared to the reference Dayton EMM-6 (red). The Knowles response is very close to what's stated in their data sheet. The thinner, dashed red curve is the Dayton curve shifted up to make it easier to compare to the Knowles curve.

MEMS Calibrated Microphones



Cursors

AX: 1.007ms
AY: 450mV
BX: 1.052ms
BY: -153.3mV
ΔX: 45us
ΔY: -603.3mV
1/ΔX: 22.22kHz

VRMS1 227.25mV
Freq1 21.886kHz
VRMS2 *****

Scope 1 (left): the MEMS microphone picks up 22kHz sound waves just fine. According to the data sheet, it will work up to at least 80kHz. The sensitivity drops off above about 25kHz, but it will definitely still pick up signals above that.

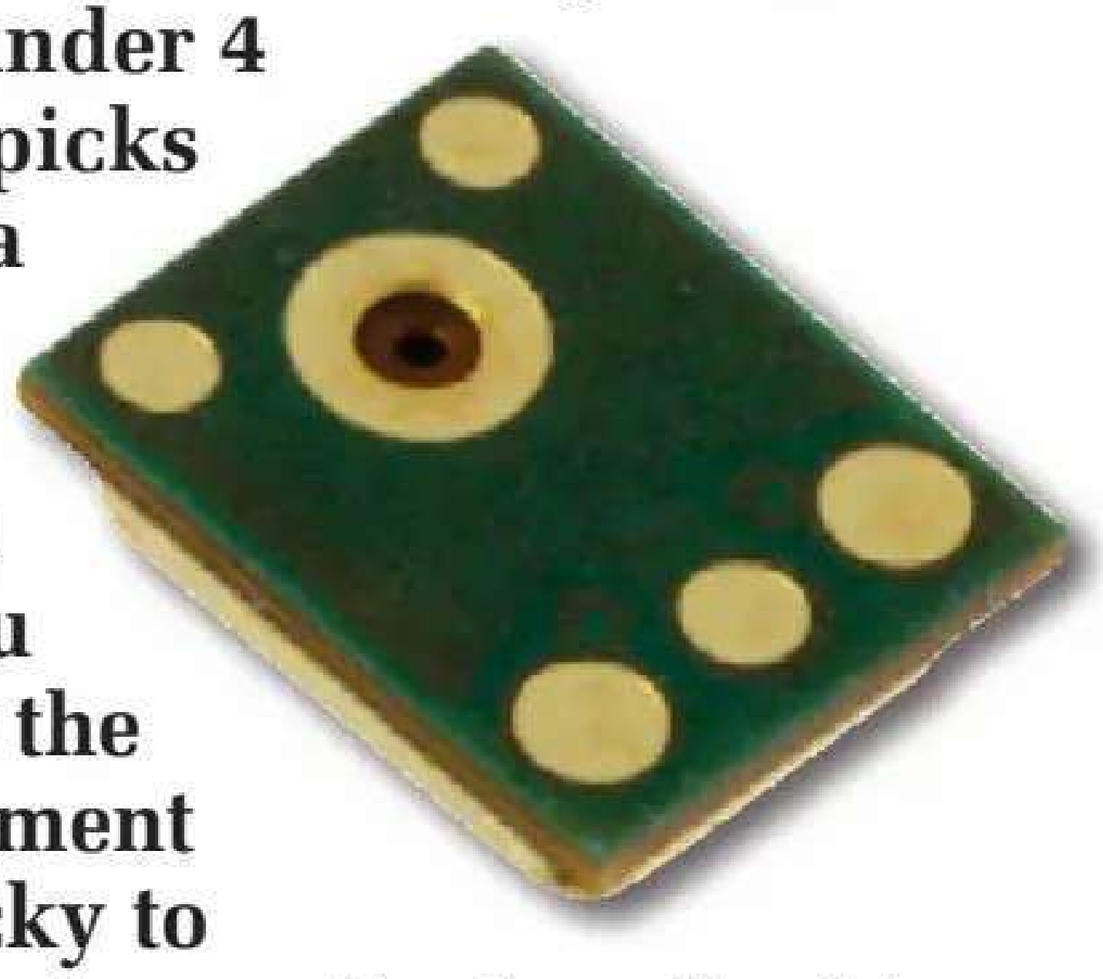


Photo 1: this MEMS microphone has a footprint under 4 × 3mm and picks up sound via the small 'acoustic port' hole in the base. You can see how the pad arrangement makes it tricky to solder; the only practical method is reflow (IR or hot air).

file or make one by taking data from the published curves.

In my opinion, the critical frequency response areas are in your crossover zones, typically in the 100-5000Hz region, making these microphones an interesting option if you are OK fiddling with tiny ICs.

Richard was interested in using them to measure the output of ultrasonic parking sensors. The only ultrasonic source I knew I had was an old-school remote from the 1960s, in which the 'buttons' make spring-loaded hammers tap brass rods. The resulting ultrasonic signals were picked up by the TV set. It was an unusual arrangement!

I used this circuit to measure the output of that remote control, with the result shown in Scope 1. The two buttons generate high frequencies at relatively high levels; the one shown in Scope 1 is at 22kHz. That is above the range of human hearing, although it might freak out your dog or cat! The bursts are short, so if you could hear them, it would be as a click.

So, as far as I can see, these are a real option for ultrasonic measurements. They are also pretty good for use as a basic calibrated microphone over the audible frequency range. **PE**

Parts List – MEMS Reference Microphone

SMD version

1 double-sided PCB coded 01108231, 64 × 13mm
1 Knowles SPU0410LR5H MEMS microphone on carrier PCB

Semiconductors
2 BC860 45V 100mA PNP transistors, SOT-23 (Q1, Q2)
1 BC849C 30V 100mA NPN transistor, SOT-23 (Q3)
3 6.8V ¼W zener diodes, SOT-23 (ZD1-ZD3) **[BZX84C6V8]**
1 3.3V 0.6-1W axial zener diode (ZD4) **[1N4728]**

Capacitors *(M2012/0805 50V X7R, unless otherwise noted)*
1 100µF 50V radial electrolytic (max 8mm diameter)
1 100µF 10V low-ESR radial electrolytic
1 10µF 16V X5R
3 1µF 50V non-polarised SMD electrolytics, no more than 4mm in diameter
2 2.2nF 5% NP0/C0G
2 1nF 5% NP0/C0G
2 470pF 5% NP0/C0G

Resistors *(all SMD M2012/0805 size 1%, unless noted)*
2 150kΩ 1 100kΩ 1 39kΩ 1 5.6kΩ
1 2.2kΩ 1 1kΩ 1 330Ω 2 47Ω
1 3.3kΩ (through-hole or SMD, ¼W 1%)

Through-hole version

1 double-sided PCB coded 01108232, 99 × 13mm
1 Knowles SPU0410LR5H MEMS microphone on carrier PCB

Semiconductors
2 BC560 45V 100mA PNP transistors, TO-92 (Q1, Q2)
1 BC549C 30V 100mA NPN transistor, TO-92 (Q3)
3 6.8V 400mW or 1W axial zener diodes (ZD1-ZD3) **[1N754]**
1 3.3V 0.6-1W axial zener diode (ZD4) **[1N4728]**

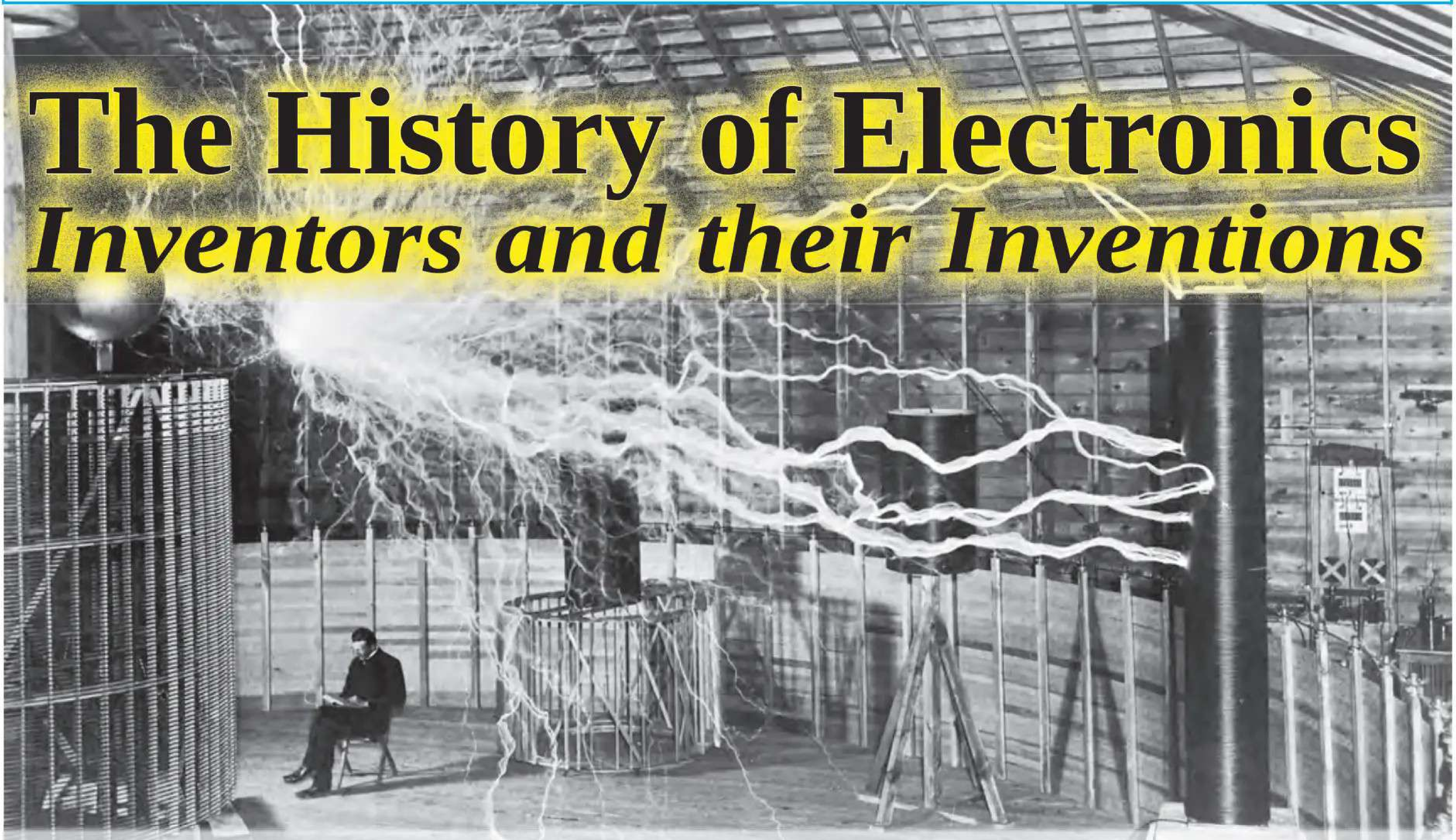
Capacitors
1 100µF 50V radial electrolytic (maximum 8mm diameter)
1 100µF 10V low-ESR radial electrolytic
1 10µF 35V radial electrolytic
3 1µF 63V/100V MKT
2 2.2nF 63V/100V MKT
2 1nF 63V/100V MKT
2 470pF 50V C0G/NP0 ceramic

Resistors *(all axial ¼W 1%)*
2 150kΩ 1 100kΩ 1 39kΩ 1 5.6kΩ
1 3.3kΩ 1 2.2kΩ 1 1kΩ 1 330Ω
2 47Ω

This is an updated version of the parts list from the September 2024 issue. In short, the changes were the addition of the SPU0410LR5H MEMS microphone, 3.3V zener diode, 3.3kΩ resistor; and the removal of one each of the 10kΩ and 2.2kΩ resistors. The case parts are not included; see the September 2024 issue for those.

The History of Electronics

Inventors and their Inventions



Physicist Isaac Newton wrote, "If I have seen further it is by standing on the shoulders of Giants". The field of electronics is no different; we could not have the technology we have today without the contributions of thousands of brilliant people. This series of articles is about them.

Part 4: by Dr David Maddison

The first three articles of this six-part series, published over the last few months, listed significant electronics-related inventions of individual inventors who were born before 1882.

This part will cover individual inventors born from 1882 to 1969. For the following two months, we will switch to discussing significant inventions credited to companies and other organisations. That will complete this series, covering inventions up to early this century.

Inventors by date of birth (1882 to 1969):

Julius Edgar Lilienfeld *1882-1963*
field-effect transistor (FET)

Lilienfeld filed for US patent 1,745,175 in 1926, which was awarded in 1930, for the field-effect transistor (FET). However, he never built a practical device because of the unavailability of high-purity semiconductor materials at the time.

Max Dieckmann *1882-1960*
video camera tube "image dissector"

Dieckmann and his student Rudolf Hell (1901-2002) obtained a patent in 1927 (applied for 1925) for a video camera tube called the "image dissector". However, Philo T. Farnsworth was the first to make it actually work (see his entry on page 26).

Alexander Meissner *1883-1958*
radio navigation systems, Meissner oscillator

Invented the Telefunken Kompass Sender in 1911, one of the earliest radio navigation systems, comprising a directional beacon used to navigate Zeppelin airships (see pemag.au/link/abnm). In 1913, he discovered positive feedback as applied to vacuum tube amplifiers.

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The History of Electronics, part 4

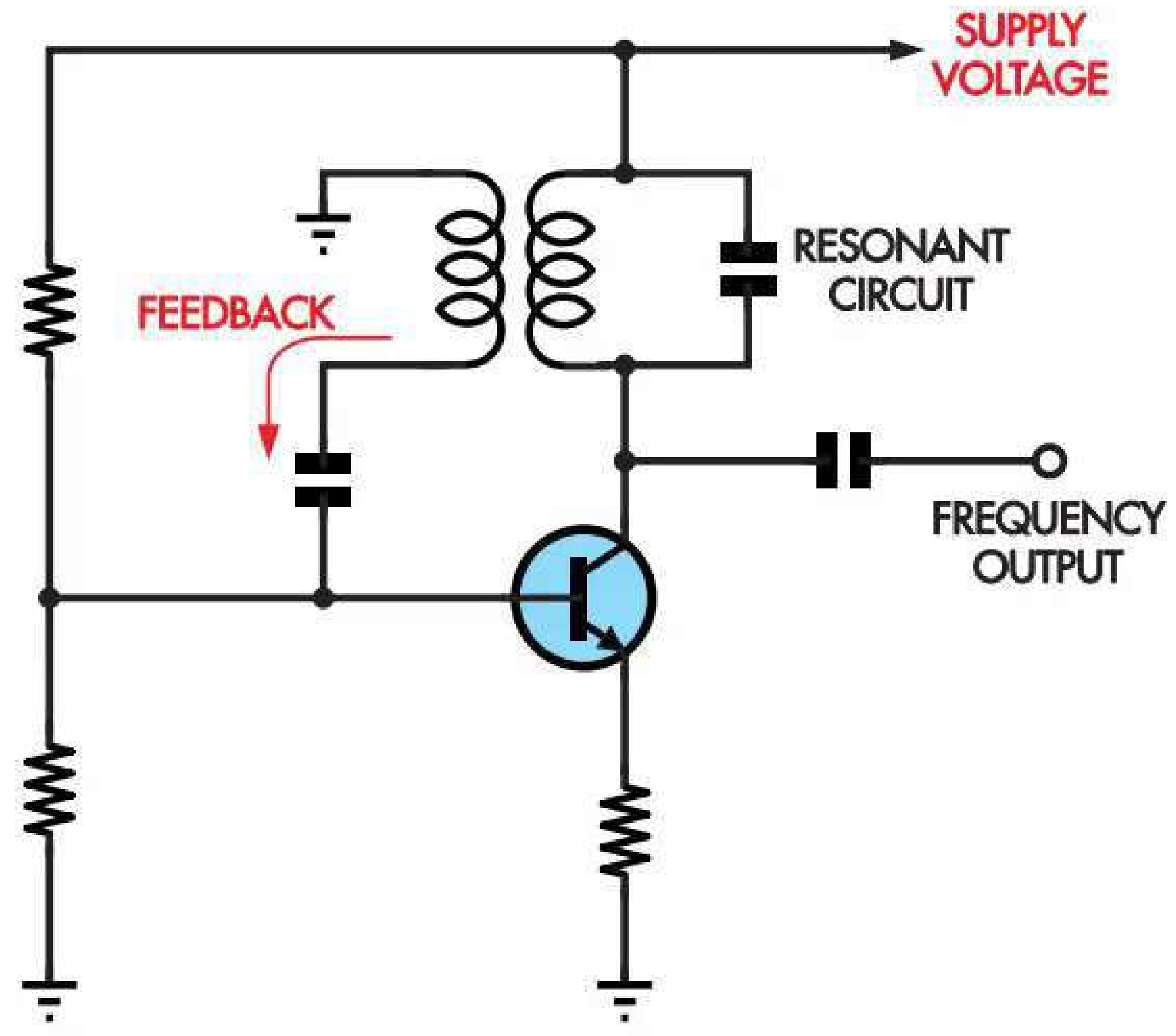


Fig.43: an Armstrong or Meissner Oscillator. *Original source:* *www.itwissen.info/en/Meissner-oscillator-127183.html#gsc.tab=0*

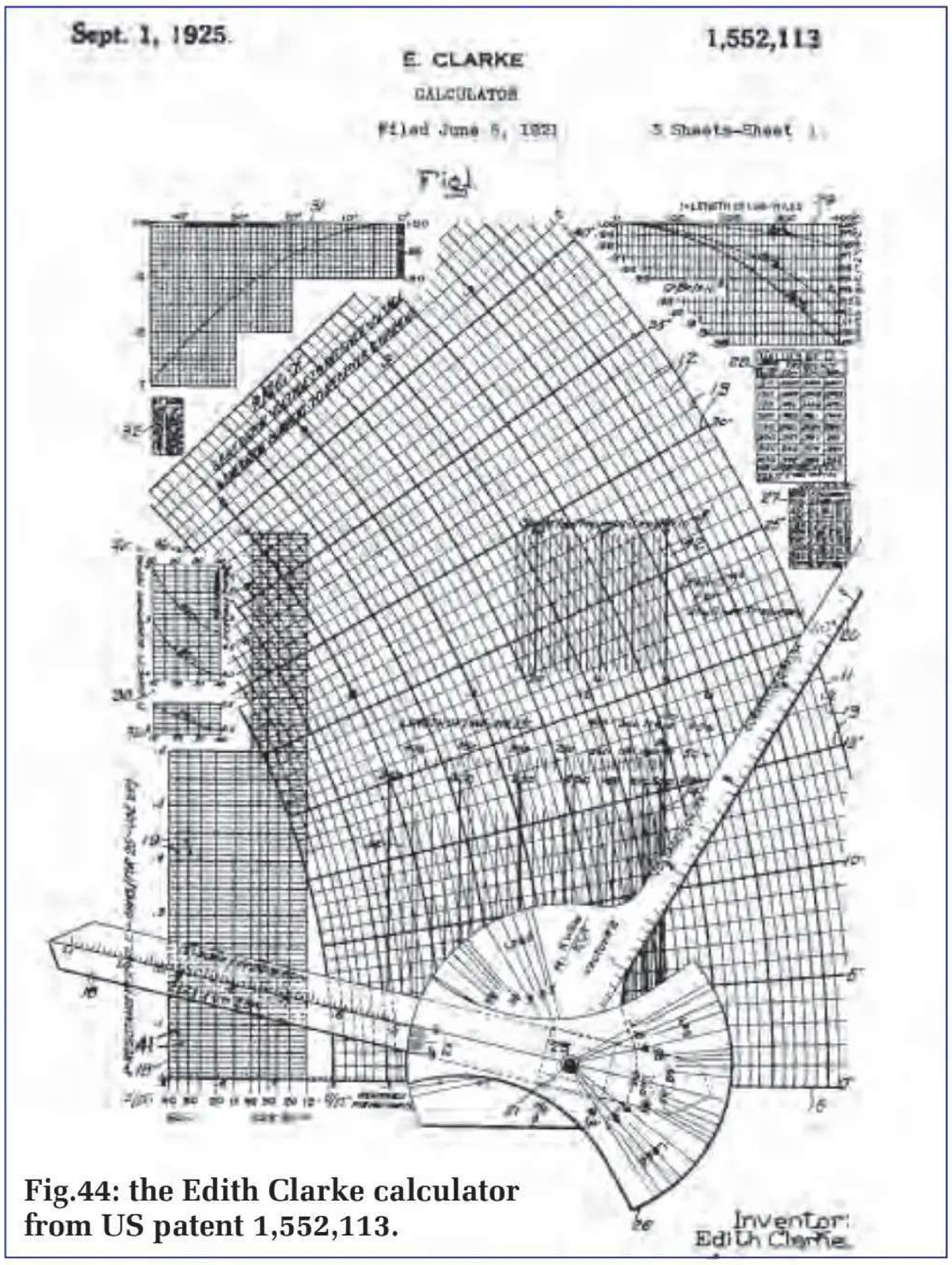


Fig.44: the Edith Clarke calculator from US patent 1,552,113.

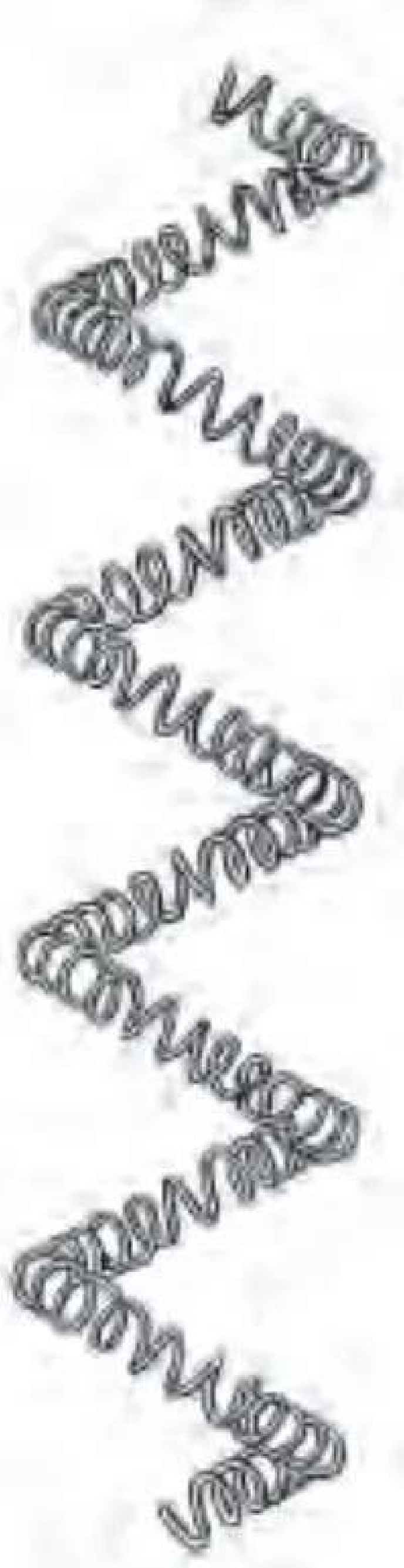


Fig.45: Burnie Lee Benbow's "coiled-coil" tungsten lamp filament from his 1917 US patent.

He co-invented the oscillator in 1913 (independently with Edwin Armstrong, 1912) and received a patent in 1920. The Armstrong oscillator or Meissner oscillator (Fig.43) uses an inductor and capacitor to produce oscillation with a valve (or transistor in modern implementations) as the amplifier. Its frequency is determined by a resonant circuit, with oscillation maintained by a feedback process.

Saul Dushman *1883-1954*
vacuum tube diodes

While at General Electric, he produced the first vacuum tube diodes in 1915, usable as rectifiers in power supplies.

Edith Clarke *1883-1959*
Clarke (graphical) calculator

Edith filed US patent 1,552,113 for the Clarke Calculator (Fig.44) in 1921, awarded in 1925. It greatly simplified calculations for long transmission lines. It was a physically simple graphical calculator, which we assume was made out of cardboard or similar, but with some complex mathematics behind it.

It also embodied a correct understanding of how inductance and capacity are uniformly distributed in long transmission lines, contrary to assumptions made at the time.

Georges Rignoux *1882-1944*
transmitting still images

Georges Rignoux and Auguste Fournier (1864-unknown) of La Rochelle transmitted still images in Paris in 1909. They were updated every few seconds, using a sensor with an 8×8 matrix of photo-sensitive selenium cells. The resolution was enough to reproduce the English (or French) alphabet.

M. Georges Rignoux.
M. A. Fournier.

Burnie Lee Benbow *1885-1976*
coiled-coil tungsten filaments

Benbow invented "coiled-coil" tungsten filaments for incandescent lamps in 1917 (Fig.45), extending their life due to less tungsten evaporation. Although simple in principle, there were enormous practical difficulties to overcome in fabrication.

Walter Han Schottky *1886-1976*
thermionic valve, schottky diodes etc

Schottky invented the screen grid thermionic valve in 1915, co-invented the ribbon microphone and ribbon loudspeaker with Erwin Gerlach in 1924 and made many contributions to semiconductor physics.

The **schottky diode** (with a metal/semiconductor junction) is named after him.

Hidetsugu Yagi *1886-1976*
Yagi-Uda antenna

Yagi published articles to the West on the Yagi-Uda antenna (Fig.46), which was invented by his assistant, Shintaro Uda (1896-1976) in 1926.

It is a directional antenna of simple design, commonly used for TV antennas and also widely used by radio amateurs.

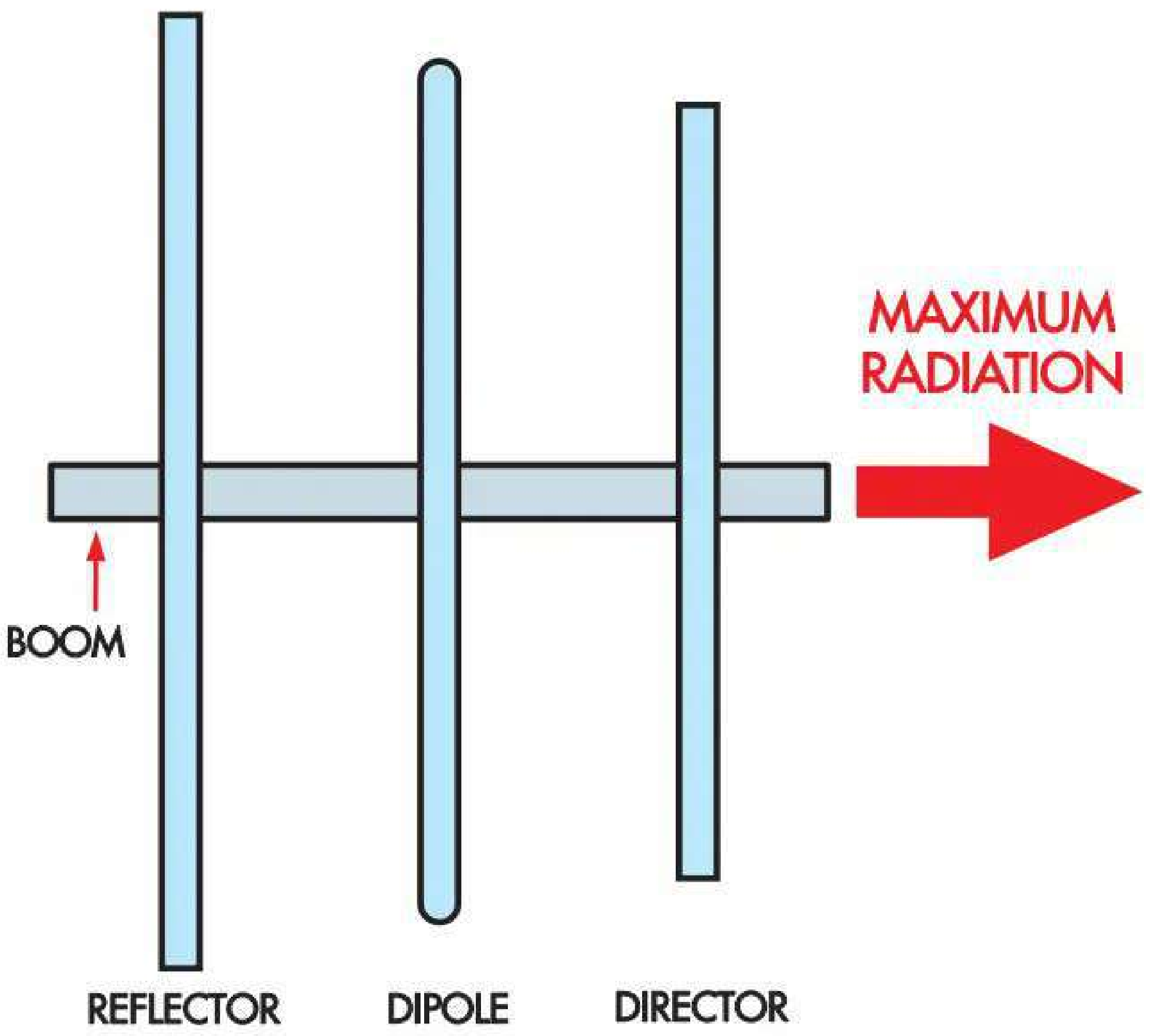


Fig.46: the basic configuration of a 3-element Yagi-Uda antenna.

John Logie Baird *1888-1946*
television

He made the first television image in 1925 (see Fig.47). It was of a rotating head, made using a Nipkow disk with 30 vertical lines of resolution. In 1926, he produced the first commercial television. In 1927, he transmitted a television picture over 705km via a telephone line.

In 1928, he transmitted a television image across the Atlantic and in 1929, the BBC transmitted the first television programs. In 1940, he started work on the first single-tube electronic colour television system, Telechrome, which was demonstrated in 1944. He also worked on Phonovision between 1926 and 1928 (more on that next month).

Sir C. V. Raman *1888-1970*
Raman effect

Sir Chandrasekhara Venkata Raman and Sir Kariamanikkam Srinivasa Krishnan (1898-1961) discovered the Raman effect in 1928. It is a form of light scattering used for analysing substances. A Raman spectrometer was used on the Mars lander Perseverance.

Vladimir Kosma Zworykin *~1888-1982*
iconoscope (television camera tube)

Filed for US patent 2,141,059 for the iconoscope in 1923 (awarded 1938). This was the first practical television camera tube and it was used for the 1936 Olympics. In Europe, it was replaced that year by the Super-Emitron and Superikonoskop. However, it remained in use in the United States until 1946, when it was replaced by the image orthicon tube.

Edwin Howard Armstrong *1890-1954*
positive feedback ("regeneration"), superhet

He was interested in how vacuum tubes work; they were not understood when the triode or "Audion" was invented by Lee de Forest in 1906. As a student, Armstrong experimented with these tubes with Professor John Harold Morecroft.

Armstrong made a breakthrough discovery in 1912 that positive feedback or "regeneration" with a triode could dramatically increase the amplification possible, allowing the use of a loudspeaker rather than headphones. He also discovered that an Audion with sufficient feedback could be used to generate a high-frequency signal for radio transmitters.

A complicated 25-year legal battle ensued between him and de Forest about patent rights for these discoveries, but Armstrong retains credit. In 1918, he invented the supersonic heterodyne or superhet circuit, which enabled radio receivers to be more selective and sensitive. That invention was also subject to legal disputation with Lucien Lévy of France, with most claims awarded to Lévy.

He developed wideband FM radio and first presented a paper on the subject in 1935, published in 1936.

Imre Bródy *1891-1944*
krypton light globes

Filled light globes with krypton instead of argon in 1930, resulting in a much longer-lasting globe, becoming one of Hungary's biggest exports. The gas was expensive, so in 1937, he devised a cheaper way to extract it from the air.

Lucien Lévy *1892-1965*
superheterodyne (superhet) circuit etc

Developed a low-frequency amplifier to listen to enemy telephone communications and for other applications during WW1 (1914-1918). He invented the superheterodyne circuit, filing a patent in 1917, resulting in a patent dispute with Armstrong, resolved mostly in favour of Lévy.

Robert Watson-Watt *1892-1973*
radar

Watson-Watt worked on detecting the direction of lightning strikes to warn pilots of storms from 1916. From 1935, he started working on and developing concepts to detect aircraft using radio reflections or radar.

By the start of WW2, 19 radar stations had been established, ready for the Battle of Britain, and 50 were in place by the war's end.

Sir Edward Victor Appleton *1892-1965*
proving the existence of the ionosphere

Appleton proved the existence of the ionosphere in 1924, a layer of the atmosphere that reflects radio waves, and won a Nobel Prize for the discovery in 1947.

Homer W. Dudley *1896-1980*
Vocoder (Voice Coder) – speech analysis

He invented the Vocoder (Voice Coder) in 1936 at Bell Labs. It is a speech analysis and synthesis system to encode speech by analysing it and reducing it to a series of control signals. Those signals could be transmitted over connection with limited bandwidth, such as an undersea cable or radio link, then reconstructed to the original speech.

Based on that work, in 1937, he and Robert Riesz invented the world's first electronic speech synthesiser, the Voder (Voice Operation Demonstrator), receiving US patent 2,121,142. It had a human operator pressing keys to produce the sound and was challenging to operate. It was demonstrated at the New York World's Fair in 1939.

See the video titled "The Voder – Homer Dudley (Bell Labs) 1939" at https://youtu.be/5hyI_dM5cGo and the free eBook PDF at pemag.au/link/abnn

During WW2, he worked with Alan Turing (see his entry on page 27) on SIGSALY, a high-level cryptographic machine for voice transmissions that employed technology from Vocoder and Voder (Fig.49).

Harold Stephen Black *1898-1983*
negative feedback amplifiers, op amps

Invented the negative feedback amplifier in 1927. It increased circuit stability, improved linearity (reducing distortion), increased the input impedance, decreased the output impedance, reduced noise, enhanced bandwidth and frequency response.

Early practical applications were the reduction of overcrowding on long-distance telephone lines, improved fire control systems in WW2,

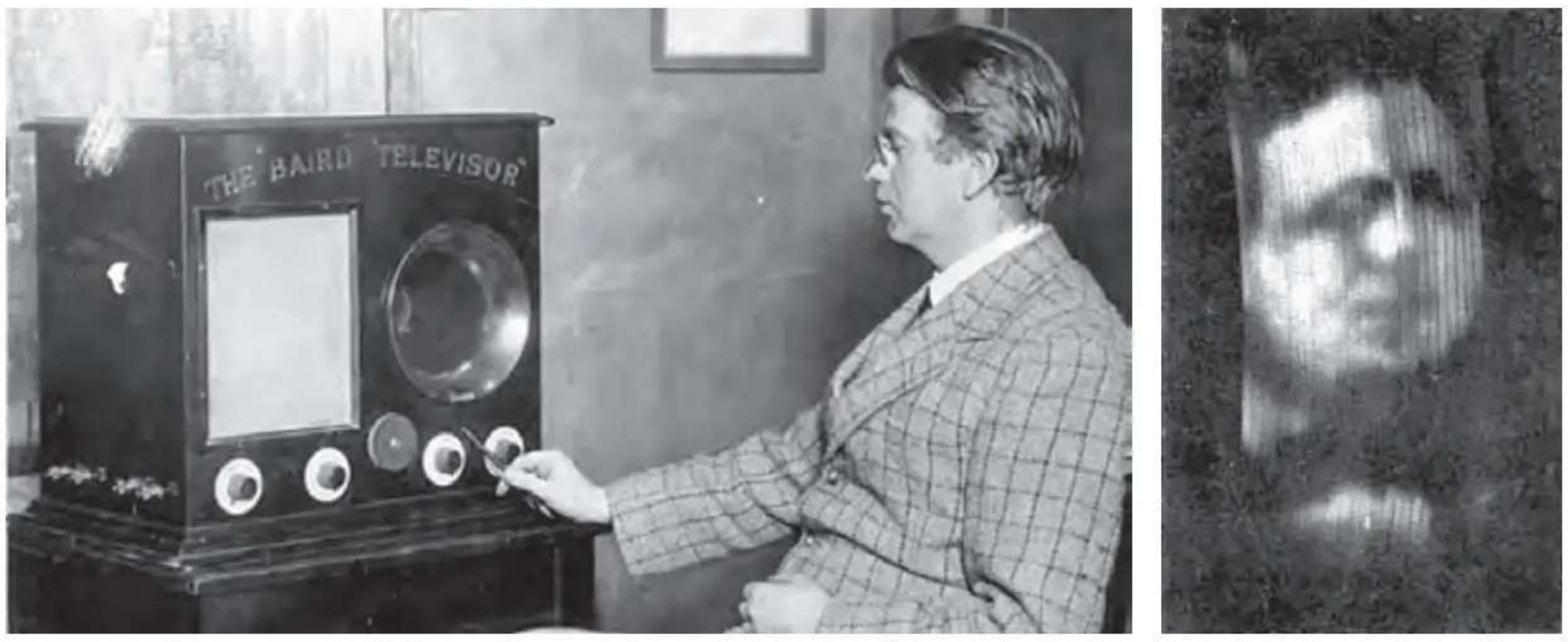


Fig.47: shown at left is John Logie Baird with his Televisor, the first commercial television from 1926. The adjacent image is of Baird's business partner, as seen on the Televisor. ***Source:*** **https://rts.org.uk/article/remembering-logie-baird-ninety-years**

The History of Electronics, part 4

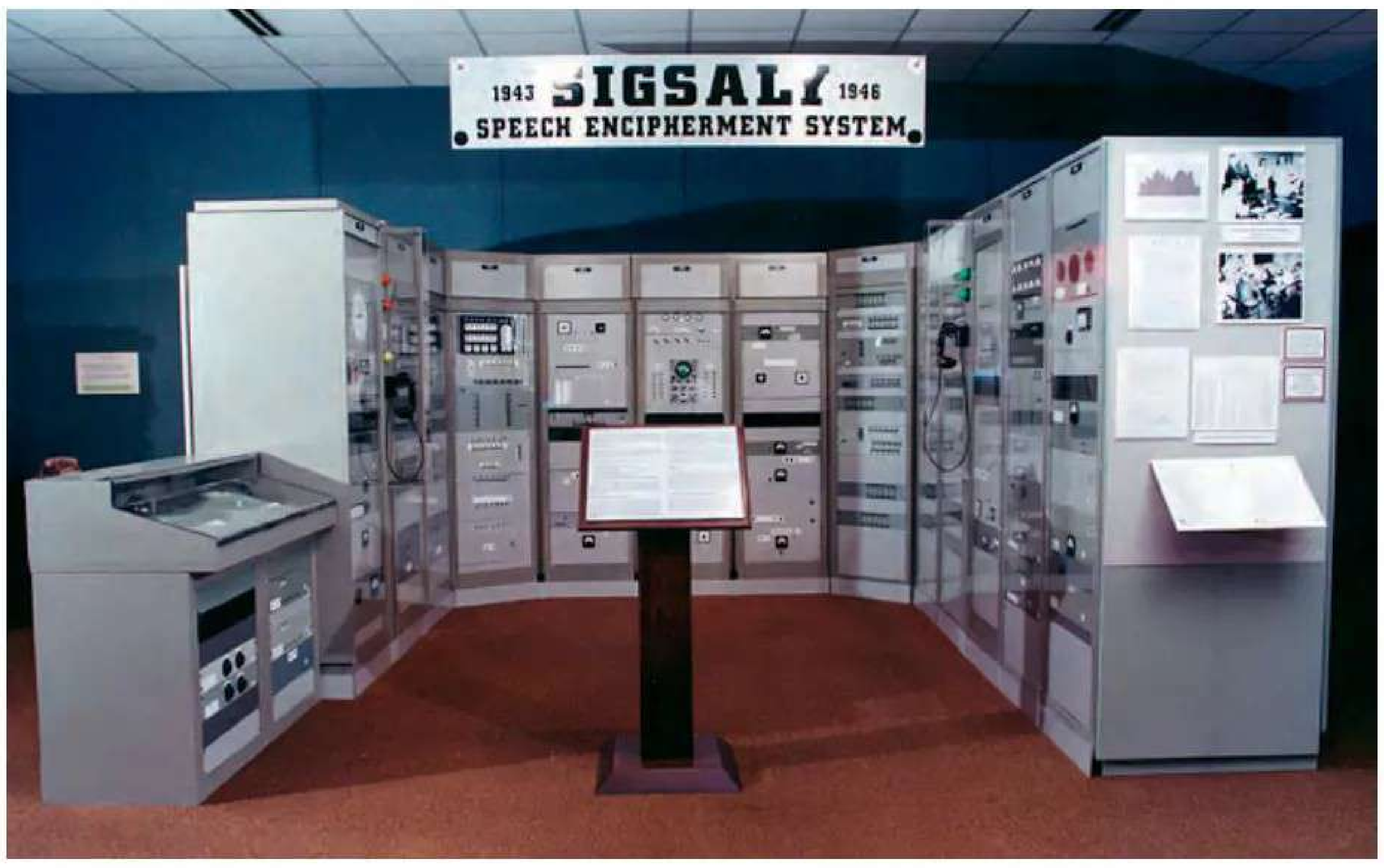


Fig.49: the SIGSALY high-level voice encryption machine used in WW2. ***Source:*** **https://w.wiki/7DAh**

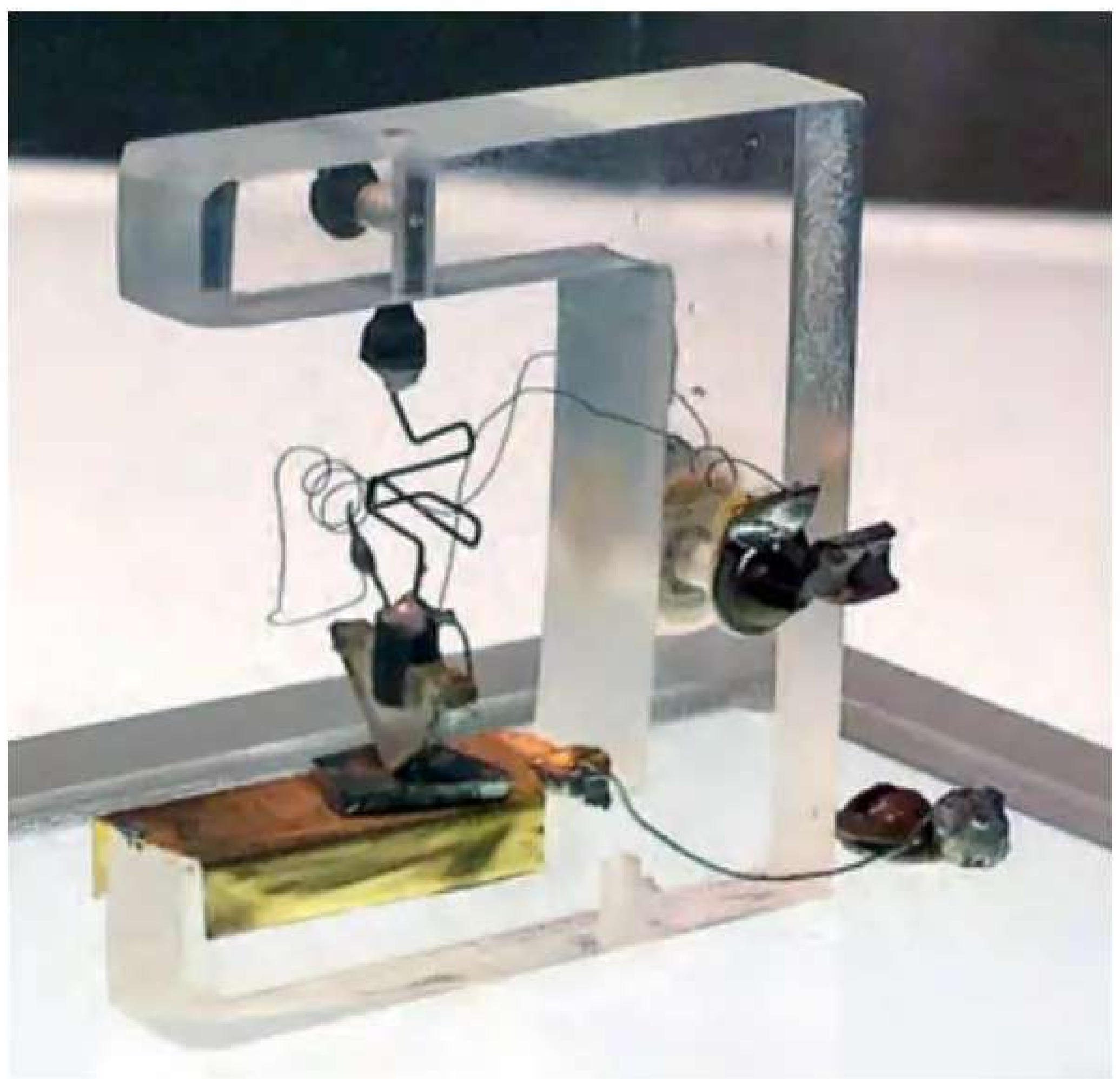


Fig.50: the first point-contact transistor from 1947. ***Source:*** **https://w.wiki/7DAi *(CC-BY-SA-3.0).***

forming the basis of operational amplifiers (op amps) and precision audio oscillators. See our article on the History of Op Amps in the previous (March 2025) issue.

Russell Shoemaker Ohl *1898-1987*
solar cell

Ohl filed for US patent 2,402,662 in 1941 for what is regarded as the world's first solar cell made with a silicon P/N junction. This design continued to be developed, reaching an efficiency of around 5% in the 1950s and 1960s.

Russell Harrison Varian *1898-1959*
klystron (linear-beam vacuum tube)

He and his brother Sigurd Fergus Varian (1901-1961) invented the klystron in 1937 and published the results in 1939. It is a vacuum tube that generates microwave frequency signals. It was the first device to generate these frequencies at a reasonable power level.

The Axis powers used it for jamming H2S radar during WW2 (many of the principles had already been published before the war). German radar used more conventional techniques to generate lower-frequency microwaves, while the Allies used the more powerful cavity magnetron (see the entry for Randall and Boot on page 26).

Kenjiro Takayanagi *1899-1990*
all-electronic television receiver

He developed the world's first all-electronic television receiver in 1926, with 40 lines of resolution. A Nipkow disc was used to scan the image at the source, but unlike other systems at the time, the receiver used a cathode ray tube to display the image.

This was months before Philo Farnsworth demonstrated the first fully electronic TV system that did not require a Nipkow disc. In 1927, Takayanagi increased the resolution to 100 lines.

Howard Aiken *1900-1973*
Harvard Mark 1

Aiken created the concept for the Harvard Mark 1, one of the earliest computers (see Fig.48). He went to IBM for funding the creation of the design, which was approved in 1939 and finished in 1944.

Dennis Gabor *1900-1979*
holography

Invented holography in 1948, a process best known for the ability to reproduce 3D images but with many other applications. He received the Nobel Prize for this work in 1971.

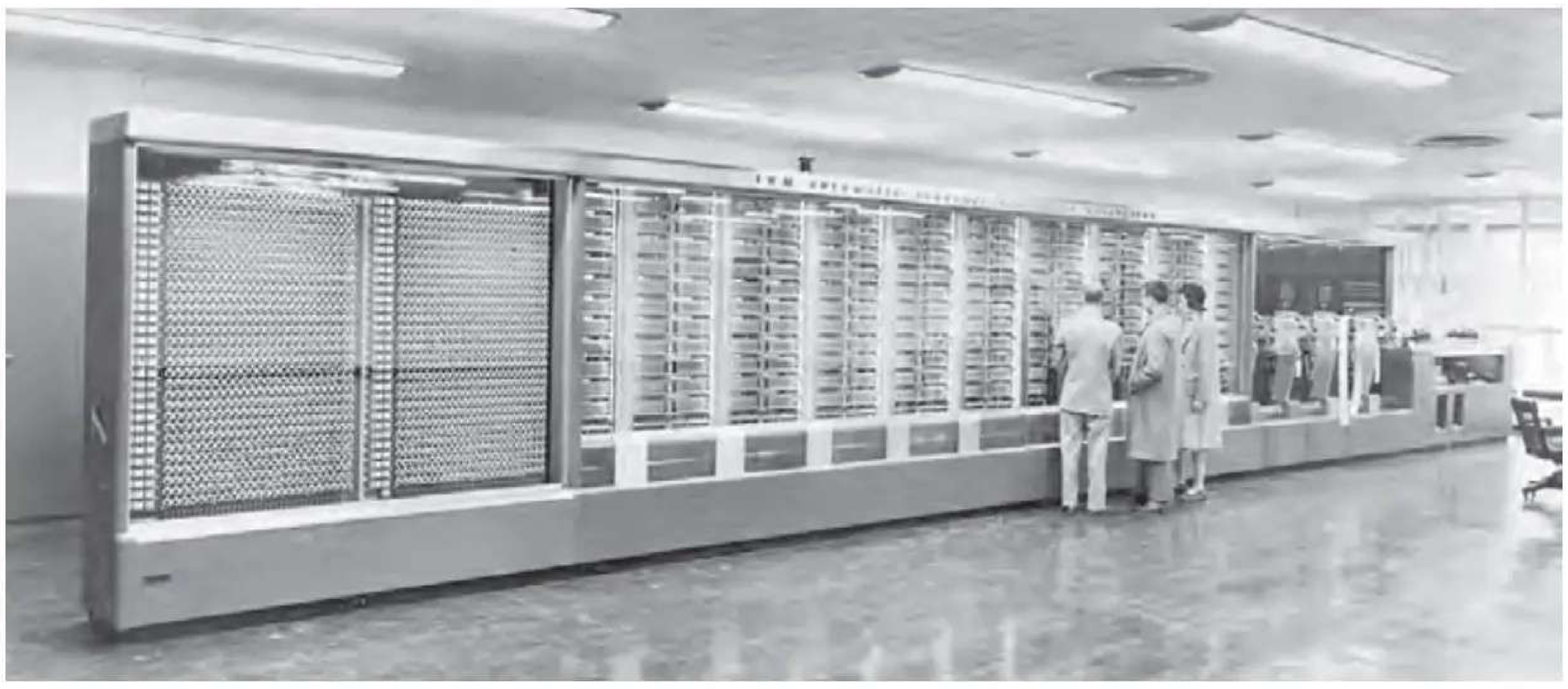


Fig.48: the Harvard Mark 1, designed by Howard Aiken, is an electromechanical computer, more than 15m long. ***Source: Encyclopædia Britannica –*** ***www.britannica.com/technology/minicomputer#/media/1/44895/19205***

Enrico Fermi & Paul Dirac
Fermi-Dirac statistics

Enrico Fermi (1901-1954) and Paul Adrien Maurice Dirac (1902-1984) independently created Fermi-Dirac statistics in 1926, which describe the behaviour of semiconductors.

Stuart William Seeley *1901-1978*
Foster-Seeley FM discriminator

Seeley and Dudley E. Foster invented the Foster-Seeley FM discriminator in 1936 and published it in 1937. It would be called a demodulator today. It reduced the cost of FM radios to a comparable level to AM radios. It was widely used until the 1970s, when ICs allowed other modulator types to be used.

Alfred Kastler *1902-1984*
optical pumping

Invented optical pumping in the early 1950s, a technique that led to the development of masers and lasers. The coherent light from lasers is crucial to semiconductor fabrication.

Walter Houser Brattain *1902-1987*
magnetometers

He worked with a group developing magnetometers during WW2 to detect submarines and applied for US patent 2,605,072 with others, including Norman E. Klein, in 1944. In 1947, with John Bardeen and William Bradford Shockley Jr, he demonstrated the first working transistor (a point-contact design) – see Fig.50.

Bardeen and Brattain were awarded a Nobel Prize for the point-contact device and Shockley for the junction transistor. Bell Labs credits 12 people as being involved with the invention of the transistor.

Alan Dower Blumlein *1903-1942*
weighting networks, stereophonic sound etc

He measured the frequency response of human ears in 1924 to design weighting networks to minimise noise and better utilise telecommunications bandwidth. In 1924, he also published work on high-frequency resistance measurements. In 1938, he submitted US patent application 2,218,902 for what was to be called an "Ultra-Linear" audio power amplifier.

In 1931, he filed UK Patent 394,325 for what is now known as stereophonic sound, but it was only commercially exploited in the 1950s after the patent expired. "Matrix processing" was used to efficiently encode sound as a common signal between left and right and a differential signal to define the spatial distribution.

After 1933, he worked on the development of television and patented several technologies, and mostly developed the 405-line Marconi-EMI TV system. During WW2, he was involved in developing the H2S radar system for the RAF to identify ground targets for night and all-weather bombing. He was killed during a flight testing the system, but it went on to be a success.

Oleg Vladimirovich Losev *1903-1942*
light-emitting diode (LED)

Extensively studied the silicon carbide point-contact junction, discovered by H. J. Round, which emitted green light. He published the results between 1924 and 1941. He produced a device, but no one saw a use for the weak light, although Losev thought it would be useful for telecommunications. We now know this device to be a light-emitting diode (LED).

John Vincent Atanasoff *1903-1995*
Atanasoff-Berry Computer (ABC)

He completed the Atanasoff-Berry Computer in 1942, which was under development since 1938. It is arguably the first digital computer, although it was not programmable, had no CPU and was not Turing complete (see Alan Turing's entry opposite).

Sir John Turton Randall *1905-1984*
cavity mangetron

Randall and Henry Albert Howard Boot (1917-1983) invented the cavity magnetron in 1940. It was an extremely important vacuum tube device used to produce high-power microwaves for radar and other applications. The klystron, as used by the Germans then, could not produce high-power microwaves. The cavity magnetron went on to be used in microwave ovens.

Philo Taylor Farnsworth *1906-1971*
fully-electronic television system

Demonstrated a fully electronic TV system in 1927 (camera and receiver). He used a video camera tube he developed, which he called the image dissector, to capture the image. He demonstrated it to the press in 1928.

Paul Eisler *1907-1992*
printer circuit board (PCB)

Eisler invented the modern printed circuit board (PCB) in 1936 while working in the UK. He had experience in the printing industry, which helped with the project. The 'intellectual property' of the invention was not well protected, as he did not read a contract he signed.

There were contributions to ideas and technologies leading up to this, such as from Thomas Edison, who made electrical tracks of glue and charcoal on a substrate in 1904; Arthur Berry, who in 1913 etched metal away to make items such as heating elements; and Charles Ducas, who described plating of copper patterns onto an insulating substrate in 1925.

Victor Ivanovich Shestakov *1907-1987*
switching circuit theory

Developed a way to implement Boolean algebra logic in electromechanical relay circuits in 1935 (switching circuit theory). This was essential for the operation of computers and other digital devices.

Claude Shannon independently invented the same theory (see his entry opposite), as well as Akira Nakashima (1908-1970).

Manfred von Ardenne *1907-1997*
3NF vacuum tube

He obtained a patent for the 3NF vacuum tube in 1923, at age 15. It had three integrated triodes (akin to an integrated circuit) and was used in the low-cost Loewe-Ortsempfänger OE333 AM radio (Fig.52).

He also produced the flying-spot scanner as a television camera in 1930 (although not a camera tube, as such) and demonstrated it at the Berlin Radio Show in 1931.

John Bardeen *1908-1991*
point-contact transistor

Bardeen and Walter Houser Brattain demonstrated the first working point-contact transistor in 1947.

Oskar Heil *1908-1994*
microwave vacuum tube

Published a paper in 1935, along with his wife Agnessa Arsenjeva, for a microwave vacuum tube, which subsequently led to the production of the first practical device. It predated the invention of the klystron, another type of microwave vacuum tube.

He also invented the air motion transformer, used in certain high-end loudspeakers (there is a video on it at https://youtu.be/-wYxHYVO6sU).

Konrad Zuse *1910-1995*
first Turing-complete computer

He invented the first programmable "Turing-complete" computer in Germany in 1941.

William Shockley *1910-1989*
transistor

He led a research group at Bell Laboratories that included the co-inventors of the transistor, John Bardeen and Walter Houser Brattain, who produced the first transistor in 1947.

In 1956, he founded Shockley Semiconductor Laboratory in Mountain View, California, but unfortunately, he was regarded as a very poor manager.

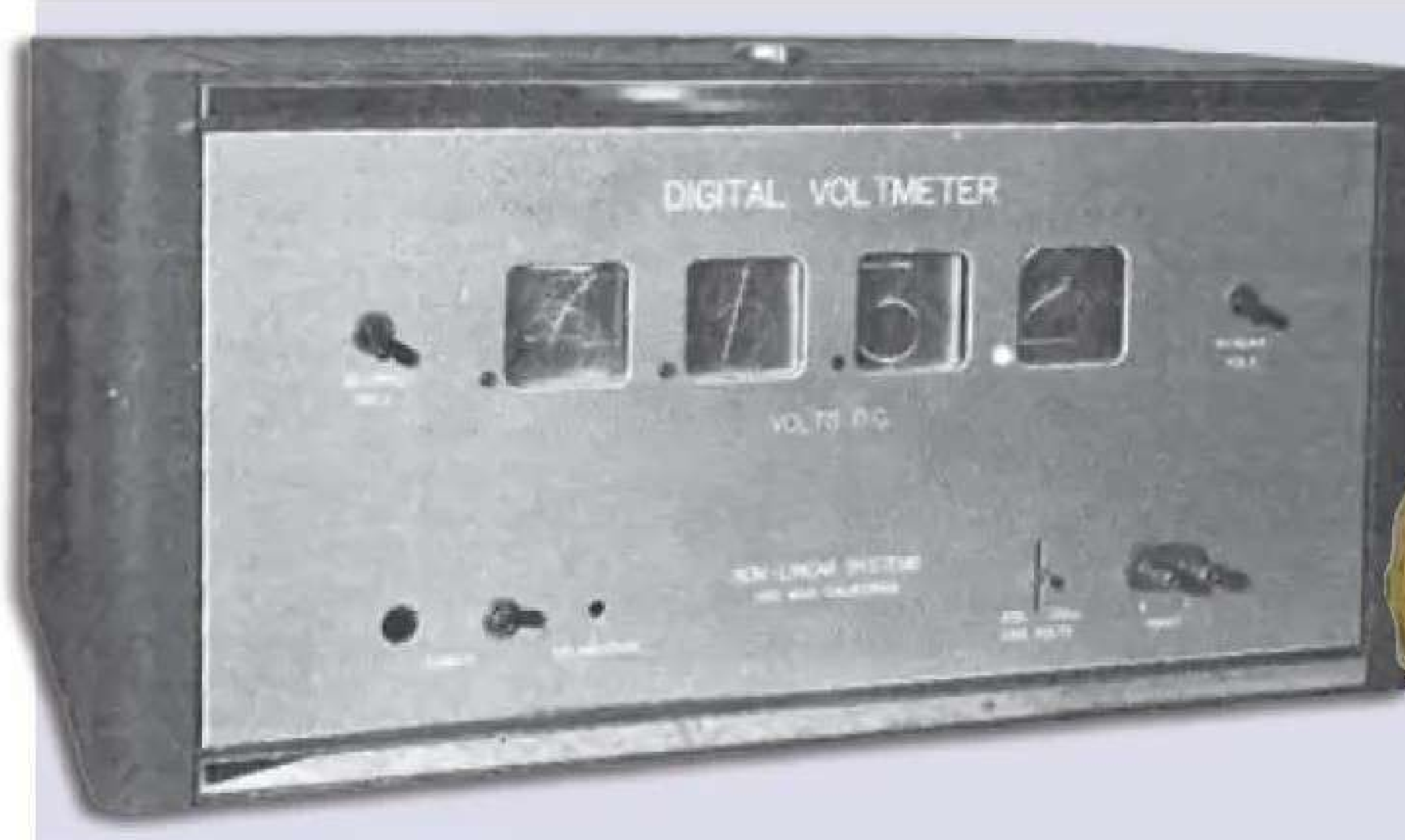


Fig.51 (left): the first digital voltmeter from 1952.

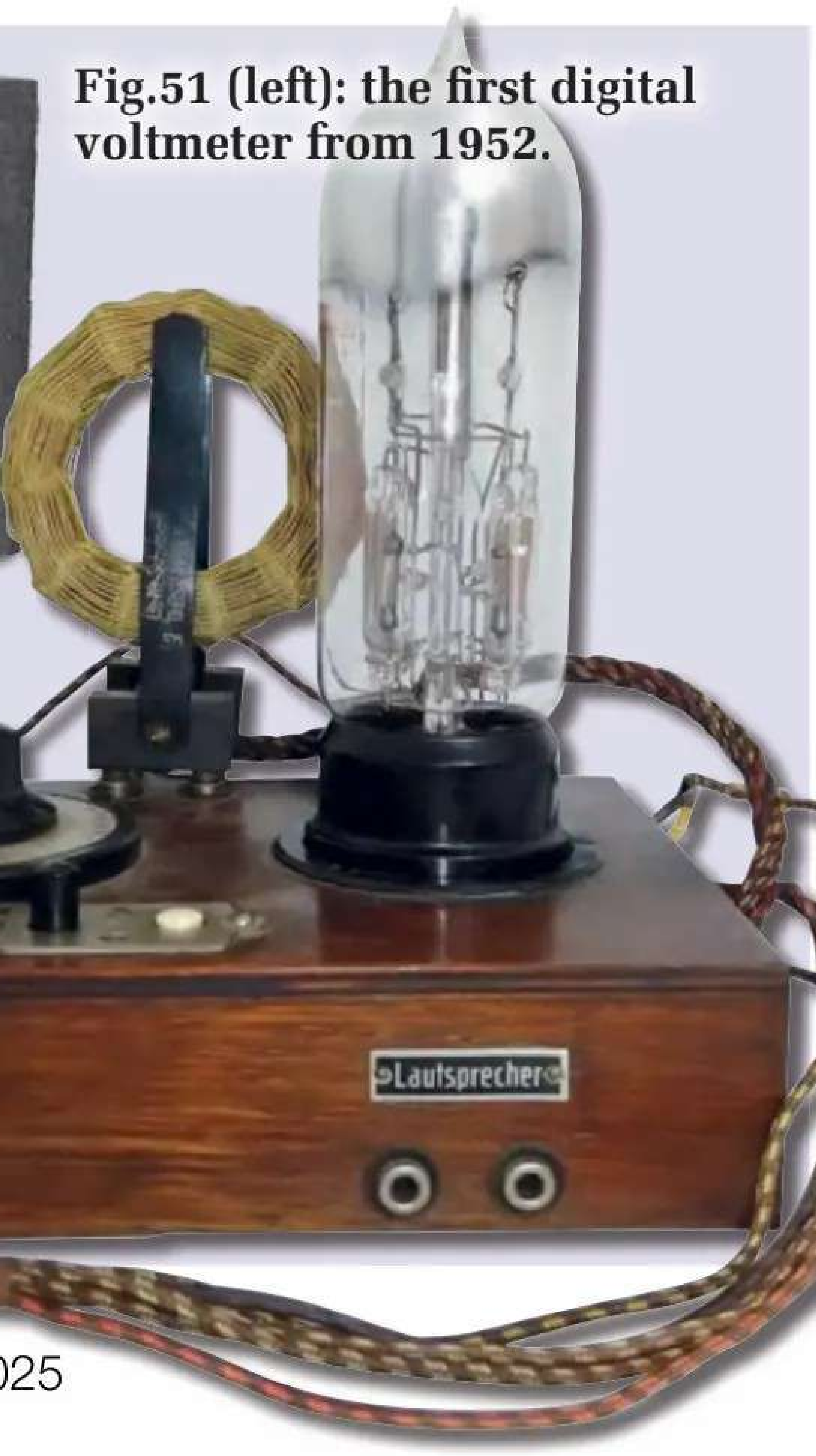


Fig.52: the Loewe-Ortsemfänger OE333 AM radio used the 3NF vacuum tube made by Manfred von Ardenne.

The History of Electronics, part 4

This led to the "traitorous eight" leaving and founding Fairchild Semiconductor in 1957. For more on this, seethearticle in the June 2022 issue of Silicon Chip magazine on on IC Fabrication (part 1; siliconchip.au/Series/382).

John Robinson Pierce 1910-2002
communications satellites

Published an article titled "Orbital Radio Relays" in the journal Jet Propulsion in April 1955. He was a pioneer of communications satellites and participated in the development of Telstar 1. Arthur C. Clarke acknowledged Pierce as one of two pioneers of such satellites, along with Harold Allen Rosen.

Hedy Lamarr 1914-2000
radio guidance system

In the early 1940s, along with George Antheil, she developed spread spectrum and frequency-hopping technology to create an unjammable (at the time) torpedo guidance system. Both techniques were used in later communications systems.

Alan Turing 1912-1954
cryptography, Turing machine etc

Turing is one of the founders of computer science and a significant figure in the development of cryptography. He created the concept of the Turing machine that can be used to compare the capabilities of different kinds of computers and the Turing test to determine if a machine can fool a human into thinking it's another human.

Claude Shannon 1916-2001
signal flow graphs, Minivac 601 computer

Demonstrated circuits in 1936 to simplify the arrangement of relays in telephone network switches. He also invented signal flow graphs in 1942.

In 1961, he designed the Minivac 601 electromechanical computer for educational purposes. There are plans to build a replica at pemag.au/link/abno

The German's Enigma machine from WWII was cracked by Alan Turing and others. ***Source:*** **https://w.wiki/7Dwg (CC-BY-SA-4.0).**

Sir Arthur Charles Clarke 1917-2008
communications satellites

He wrote a *Wireless World* article in 1945 proposing what we would now call communications satellites (in particular, geostationary satellites).

Harry Wesley Coover Jr 1917-2011
super glue

Invented cyanoacrylate adhesives ('super glue') in 1942. A commercial product was not released until 1958, marketed by Kodak as Eastman 910. These adhesives bond almost instantly and have wide application in commercial electronic assembly.

Andrew F. Kay 1919-2014
digital voltmeter

He invented the digital voltmeter (Fig.51) in 1952.

Otis Frank Boykin 1920-1982
precision wire-wound resistors, pacemakers

Produced many inventions, including an improved form of precision wirewound resistor with low inductance and reactance. He also invented a precision control unit for cardiac pacemakers in 1964.

Norman Joseph Woodland 1921-2012
barcode

He applied for a patent for a barcode in 1949, to encode price and product description and other data (see US patent 2,612,994). It was a sound idea, but there was not yet a suitable computer to implement it.

Rubin Braunstein 1922-2018
gallium and indium-based semiconductors

He measured infrared emission from devices he made from the semiconductors gallium arsenide (GaAs), gallium antimonide (GaSb) and indium phosphide (InP) in 1955. This is the basis for LED lights and semiconductor lasers.

David Paul Gregg 1923-2001
optical disc

Invented the optical disc in 1962 (although it was discussed as early as 1958). He filed for US patent 3,381,086 in 1962, granted in 1968.

Jack St. Clair Kilby 1923-2005
first integrated circuit (IC) etc

He is credited for the first integrated circuit (IC) in 1958, along with Robert N. Noyce. He also invented the handheld calculator and thermal printer.

Seymour Cray 1925-1996
CDC660 supercomputer

Cray designed the first silicon transistor supercomputer in 1964, the CDC660, considered the first successful supercomputer. Germanium transistors, in use until that time, were not fast enough. It was the fastest computer in the world at the time, about ten times faster than others.

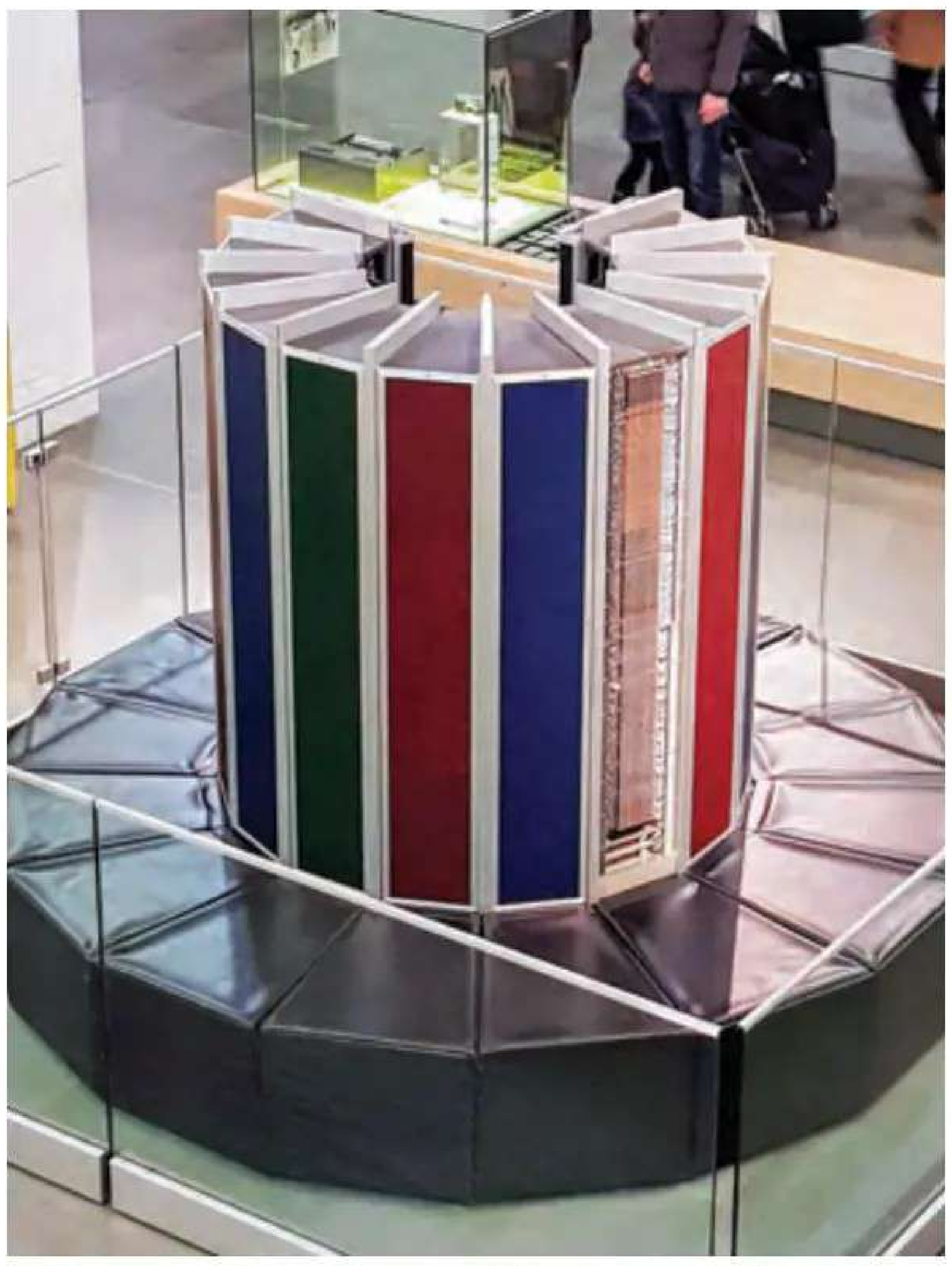


Fig.53: a Cray-1 on display at the Science Museum in London. ***Source:*** ***https://w.wiki/7DBY*** **(CC-BY-SA-2.0).**

In 1972, Cray started his own company, Cray Research, and designed the famous Cray 1 (See Fig.53). It was released in 1976 and became one of the most successful supercomputers.

Narinder Singh Kapany 1926-2020
fibre optics

Kapany invented fibre optics (he coined the term). In 1953, along with Harold Horace Hopkins (1918-1994), he transmitted an image through a bundle of 10,000 optical fibres with better image quality than had previously been achieved. This led to the first practical gastroscope for medical investigations, developed by other researchers in 1956.

Junichi Nishizawa 1926-2018
avalanche photodiode, solid-state maser etc

Nishizawa invented the avalanche photodiode in 1952, a solid-state maser in 1955 and, in 1963, proposed the idea of fibre-optic communications. He also patented graded-index optical fibres in 1964. Among his other inventions was the static induction thyristor in 1971.

Robert Norton Noyce 1927-1990
monolithic silicon IC

Noyce invented the monolithic silicon integrated circuit in 1959 and co-founded Fairchild Semiconductor in 1957 and Intel Corporation in 1968. Credit is also given to Jack Kilby for the invention of the integrated circuit.

Theodore Harold Maiman *1927-2007*
laser

Invented the first laser in 1960, a device to produce light with all emissions of the same wavelength and all in phase.

Nick Holonyak Jr *1928-2022*
visible light laser diode

He invented the visible light laser diode in 1962. It lased at low temperatures and functioned as an LED at room temperature.

Manfred Börner *1929-1996*
optical fibre communication system

Demonstrated the first working optical fibre communication system at Telefunken Research Labs in 1965.

James Robert Biard *1931-2022*
infrared LED

Biard held numerous patents and also invented, along with Gary Pittman, an infrared LED in 1961 (receiving US patent 3,293,513). In 1962, Texas Instruments released the first commercial LED (SNX-100) for US$130 each, almost $2000 today!

Chris Wallace *1933-2004*
hardware binary multipler

Wallace invented the hardware binary multiplier in 1964 for arithmetic operations in computers.

Sir Charles Kuen Kao *1933-2018*
reducing signal attenuation in optical fibres

He and George Alfred Hockham (1938-2013) at British STC proposed that making optical fibres out of more pure materials could dramatically reduce signal attenuation in 1965. Today, losses in optical fibres are extremely low, making repeaters only necessary every 70-150km.

George Harry Heilmeier *1936-2014*
liquid crystal displays (LCDs)

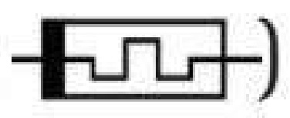
Discovered effects in liquid crystals in 1964, which led to the first liquid crystal displays (LCDs) using what he called dynamic scattering mode (DSM).

Gary Keith Starkweather *1938-2019*
laser printer

He invented the laser printer in 1969. The first commercial laser printer on the market was the IBM 3800, released in 1976 to replace line printers, with the Xerox 9700 following in 1977 for high-quality printing.

Another reason the 9700 is significant is what happened when Xerox refused to supply code for that printer (as they had done for a previous model) in 1980. That led to Richard Stallman (see below) and others at the MIT AI Lab starting the free software movement.

Richard Stanley Williams *1951~*
memristor

Developed a practical version of the memristor (memory resistor ) at HP in 2008. The memristor was first postulated in 1971 by Leon Ong Chua (1936~).

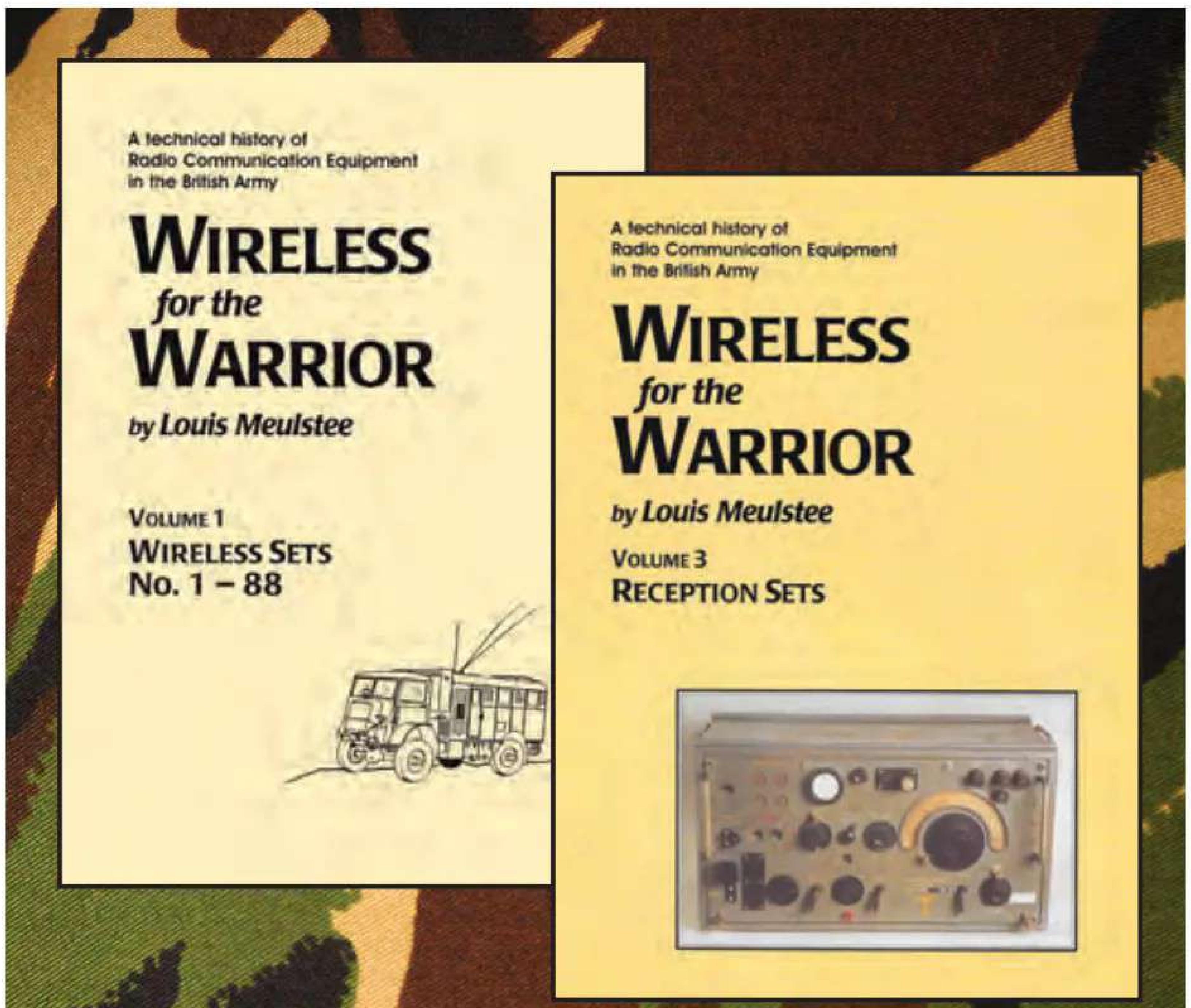
Richard Matthew Stallman *1953~*
GNU project, GCC, Emacs

He started the free software movement in 1980 and, in 1983, founded the GNU Project. He also founded the Free Software Foundation (FSF) in 1985. The tools developed by the GNU Project were instrumental for Linus Torvalds and others to make Linux a practical operating system.

Linus Torvalds *1969~*
Linux operating system

The driving force behind the open-source Linux operating system. However, thousands of others have significantly contributed to its development, including Andrew Morton, Alan Cox, Greg Koah-Hartman and Ingo Molnar. Linux is licensed under the GNU GPL.

Linux currently powers the majority of the world's top web servers. **PE**



A technical history of Radio Communication Equipment in the British Army
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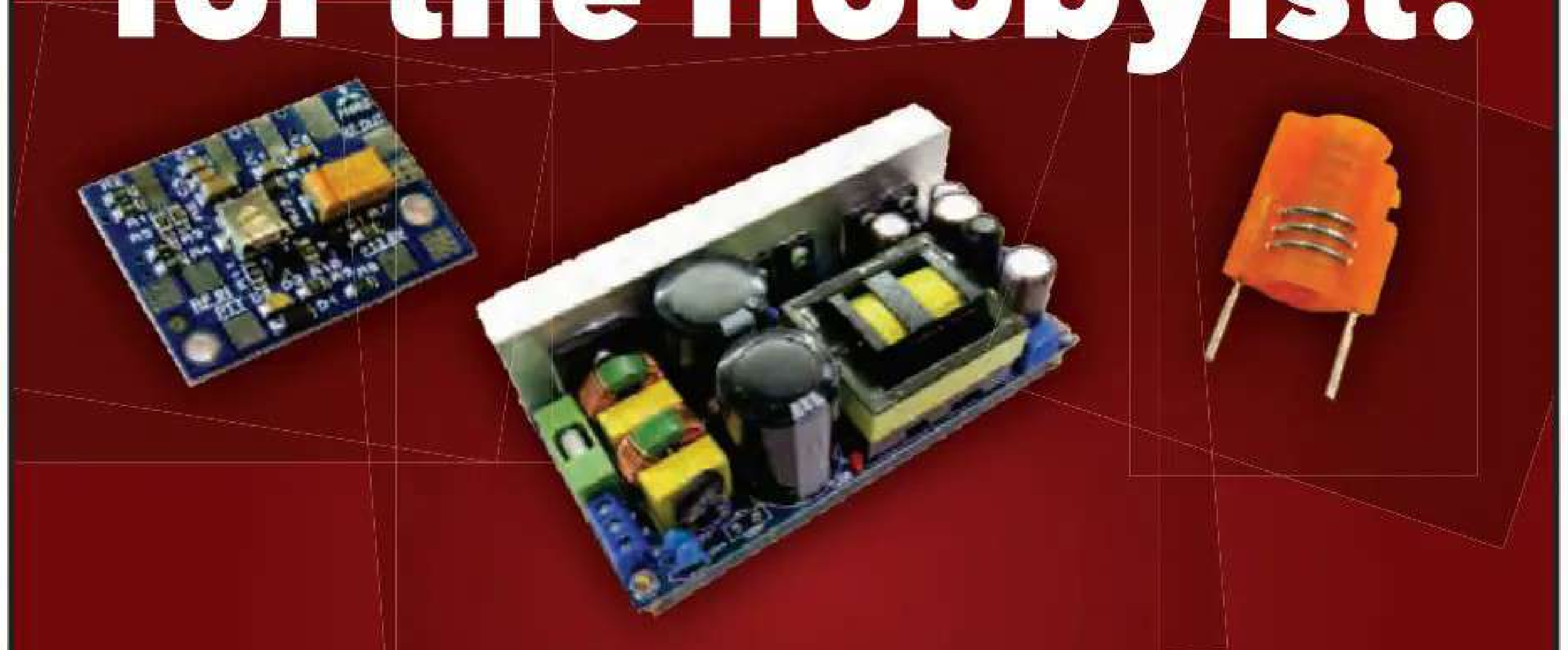

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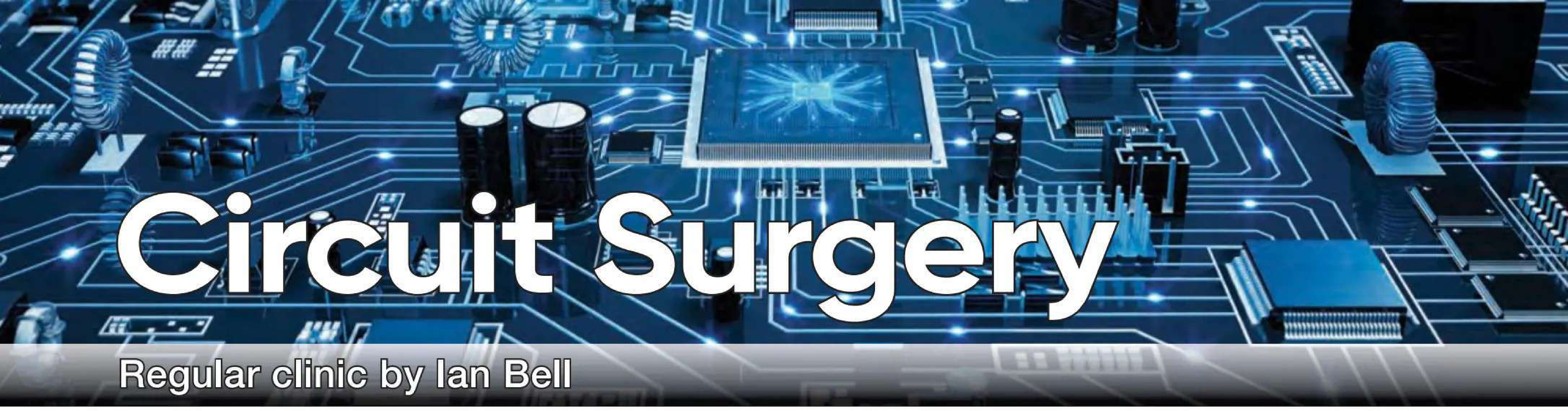


Circuit Surgery

Regular clinic by Ian Bell

Topics in digital signal processing – LTspice symbols and impulse responses

We are looking at various topics related to digital signal processing (DSP).

DSP covers a wide range of electronics applications where signals are manipulated, analysed, generated, stored or displayed as digital data but originate from and/or are converted to real-world signals for interaction with humans or other parts of the physical world.

Fig.1 shows the key elements of a generic DSP system with a signal path from an analog input via digital processing to an analog output. This does not necessarily represent every DSP system (not all have all the parts shown), but it serves as a reference for the various subsystems we will consider.

Over the last couple of months, we have discussed windowed sinc digital filters (the processing part in Fig.1) and developed LTspice simulations of these.

After initially implementing them fully using schematic drawings, we switched to text-based circuit descriptions due to the number of components (specifically behavioural sources) and signals required to realise a filter with reasonable performance.

Although there are a large number of sources and signals, the structure is repetitive, so it is reasonably straightforward to automate their creation by writing code, or partly automate it using spreadsheets.

Using text descriptions of the filter circuits facilitates defining them as LTspice subcircuits, which makes it easier to reuse them (multiple copies in the same simulation, or in a variety of different simulations). Reuse is further enhanced by creating symbols for the subcircuits, which means they can be placed on a schematic in the same way as the components provided in the download.

This month, we will describe the process of creating and using symbols in detail. Although this is in the context of using our filter simulations, user creation of symbols in LTspice is more widely applicable; for example, when downloading third-party models of devices.

Over the last couple of months, we have also discussed the concept of impulse response and its relevance to designing windowed sinc filters. This month, we also look at two methods that allow you to use LTspice to find the impulse responses of the filters to help confirm correct implementation.

Text or symbol?

Last month, we included the filter subcircuits into our LTspice simulation by adding the appropriate component statement as text on the schematic, for example:

```
X_LP1 in out_lp_rec
+ lp_5k_48k_rec_25 sp={period}
```

As discussed then, the input to the LTspice simulator is a text description of the circuit called a ‘netlist’. Netlists are automatically created from the schematic in the background to run a simulation. They can be seen using **View → Netlist** from the menu. LTspice can also open and run simulations from netlist text files – it does not need a schematic to work.

Text is placed on a schematic using the t and .t buttons in the toolbar. Any text placed on the schematic as a “SPICE directive” (.t icon), which has a default colour of black, is copied directly into the netlist. Text which is placed on the schematic as a “Comment” (t icon), which has a default colour of blue, is copied with the prefix “* ”, which designates a comment line in the netlist.

Comments are ignored by the simulator. Right-click the text on the schematic to edit it, or switch between directive and comment modes.

Using text to reference a subcircuit is straightforward, but some people may find the schematic more difficult to interpret without a symbol for the device. Fortunately, with a little bit of effort, we can create symbols for our own subcircuits and add these to schematics in the same way as standard LTspice components. Before we describe that, it is worth taking a quick look at text file handling in LTspice.

Opening LTspice text files

Although many users may only interact with LTspice via schematics and result plots in the graphical user interface (GUI), it can edit text files and has syntax colouring based on SPICE netlists (as seen in Fig.2). Commonly used files containing text circuit descriptions are netlists (.cir, .net or .sp), subcircuits (.sub) and libraries containing multiple subcircuits (.lib).

If you make frequent use of these files in LTspice, it is useful to configure your operating system to associate LTspice with the relevant file extensions. These extensions are the standard ones LTspice will open and treat as a netlist.

LTspice expects the first line of a netlist file to be a comment (text is coloured green). If you open a .lib or .sub file into LTspice directly from Windows Explorer and there is a `.subckt` statement on the first line, you may get an error message because it is treated as a comment and not parsed as a command. It is best to put a comment on the first line to avoid this problem.

With the first line as a comment, you may still get a warning about a lack of analysis requests – LTspice is trying to interpret the file as a full simulation netlist, so this can ignored for a subcircuit or library file.

If you open these files from LTspice then you will not get the errors or warnings. It will open a text file with any extension, but you have to change the file type dropdown

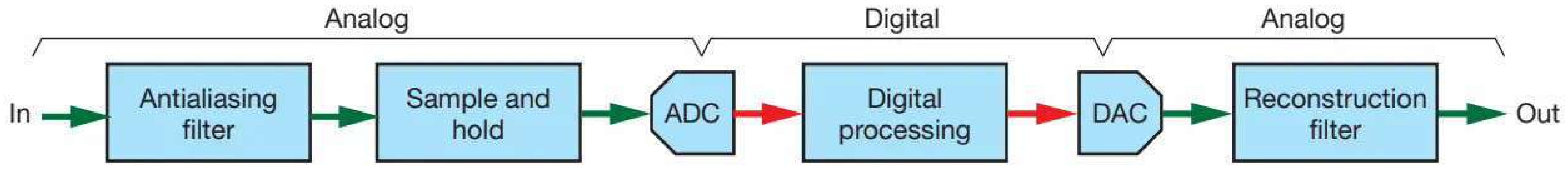
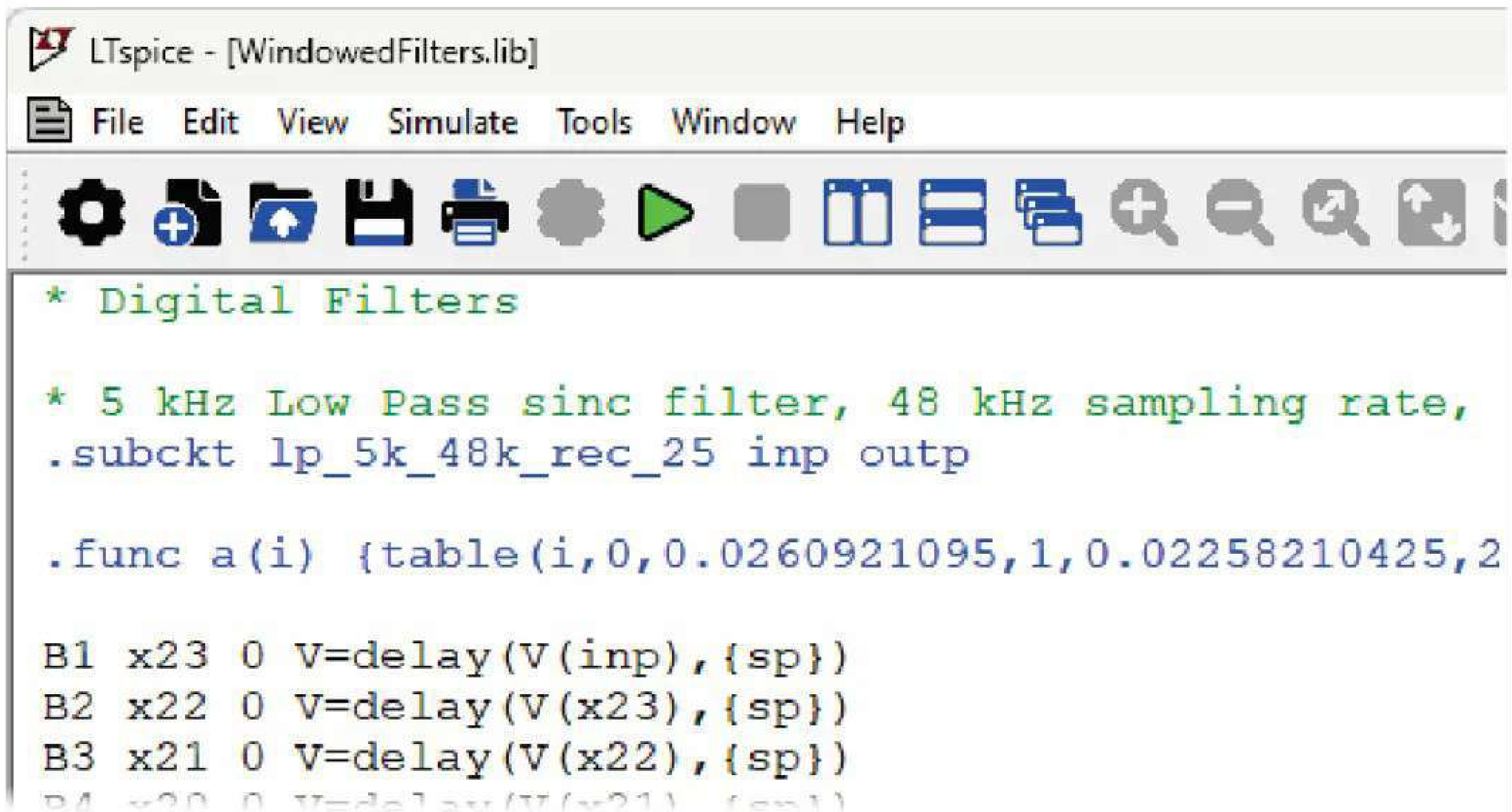


Fig.1: a generic digital signal processing (DSP) system structure.



LTspice - [WindowedFilters.lib]
File Edit View Simulate Tools Window Help

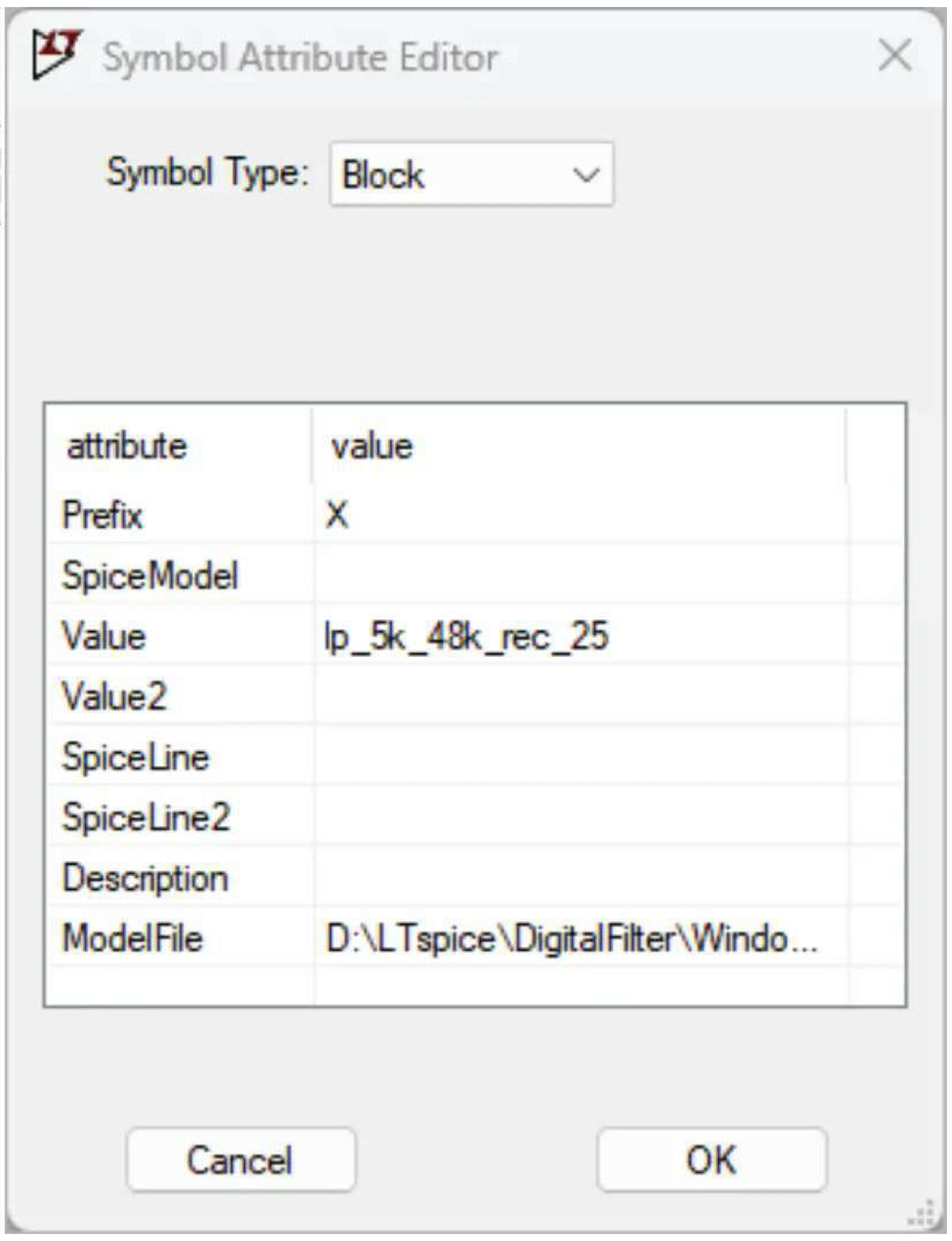
```
* Digital Filters

* 5 kHz Low Pass sinc filter, 48 kHz sampling rate,
.subckt lp_5k_48k_rec_25 inp outp

.func a(i) {table(i,0,0.0260921095,1,0.02258210425,2

B1 x23 0 V=delay(V(inp),{sp})
B2 x22 0 V=delay(V(x23),{sp})
B3 x21 0 V=delay(V(x22),{sp})
```

Fig.2: a subcircuit/library text file edited in LTspice.



Symbol Type: Block

attribute	value
Prefix	X
SpiceModel	
Value	lp_5k_48k_rec_25
Value2	
SpiceLine	
SpiceLine2	
Description	
ModelFile	D:\LTspice\DigitalFilter\Windo...

Fig.5: the LTspice symbol attribute editor dialog box.

in the file open dialog to "netlist" or "all files" to see them listed. If the file opened is a valid netlist with a simulation command (analysis request), such as **.ac.** or **.tran**, it can be simulated.

Making a symbol

LTspice has a symbol editor that allows you to draw symbols graphically and configure them to facilitate their use with the schematic editor. There are three options for creating a symbol from a schematic or subcircuit definition. You can create a symbol from scratch, auto-generate one, or copy and edit one from the LTspice library.

There are two scenarios for automatically generating a symbol. The first uses a schematic as input and is intended for use with hierarchical schematics, where schematics that are part of a larger design are included in a higher-level schematic as block symbols. As the filters we are discussing are defined by subcircuit netlists, that approach is not relevant here.

The second approach uses subcircuit definitions, which is what we need. Its primary expected use is with downloaded third-party component models, but of course, it can also be applied to subcircuits created by users.

To automatically generate a symbol from a subcircuit description, open the library/subcircuit text file in LTspice and right click on the line containing the relevant **.subckt** statement. Select "Create Symbol" from the menu and confirm creation of the subcircuit in the message box (see Fig.3). The symbol will be generated and opened in the symbol editor, as shown in Fig.4.

The filter **lp_5k_48k_rec_25** subcircuit uses a parameter (**sp**) to set the sample period; a value must be provided for this for a simulation using the subcircuit to work.

Unfortunately, automatic symbol generation does not configure the symbol to allow a parameter to be passed to it individually. When the symbol is placed on a schematic, the value can be set globally by defining a parameter with the required name, for example:

```
.param sp=20.833u
.param sp={period}
```

This is not ideal because, if multiple instances of the symbol (or other components using the same parameter) are used, they are all forced to use the same value. There are other problems with the auto-generated symbol. These can all be resolved by editing the symbol's attributes.

In the symbol editor, go to **Edit → Attributes → Edit Attributes**. This opens the attribute editor, shown in Fig.5. The exact way the attribute data is used (and the symbol behaves in the schematic editor) depends on how the attributes are specified, which can be confusing.

The most useful case for our purposes is where the "ModelFile" attribute is defined (the filename of the library/subcircuit file containing the subcircuit), which is what is auto-generated.

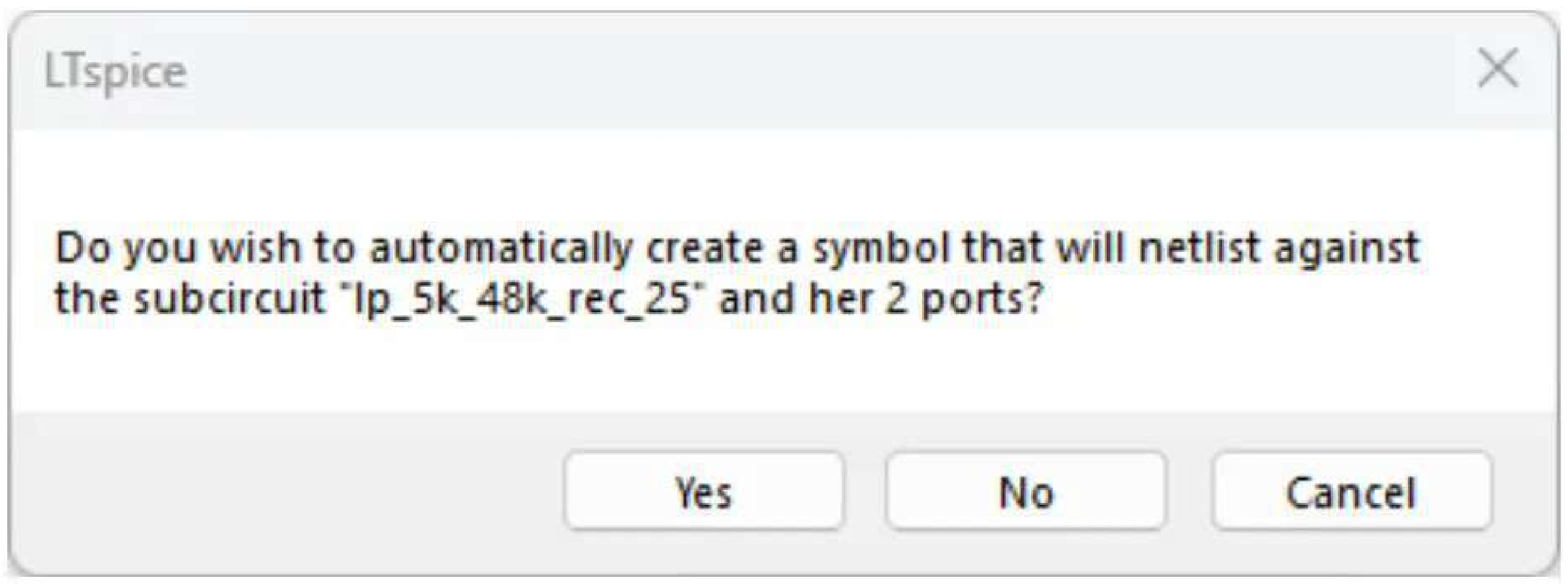
With "ModelFile" defined, the symbol is netlisted on two lines in the form shown below, and has values (such as parameters) that can be edited on the schematic where the symbol is used. In the following, **<Attrib>** refers to the value of the attribute with the name **Attrib**.

```
.lib <ModelFile>
<name> node1 node2 [...]
+ <SpiceModel> <Value> <Value2>
+ <SpiceLine> <SpiceLine2>
```

The prefix will be the first character in the name; for subcircuits, this is "X". The problems and solutions with auto-generated symbols are as follows.

- The symbol type is set to "Block", which is intended for hierarchical schematics – it should be of type "Cell".
- The ModelFile parameter is set to the full path of the library file (as seen in Fig.5). This is not ideal because, if the library files are moved, the symbol will stop working. Using just the file name means LTspice will look in the same folder as the schematic for the subcircuit definitions, wherever the schematic is.
- Parameters can be specified in the "Value2" attribute using the format **Name=DefaultValue**

Multiple parameters can be specified separated by spaces. After editing the



Do you wish to automatically create a symbol that will netlist against the subcircuit "lp_5k_48k_rec_25" and her 2 ports?

Fig.3: the confirmation message box.

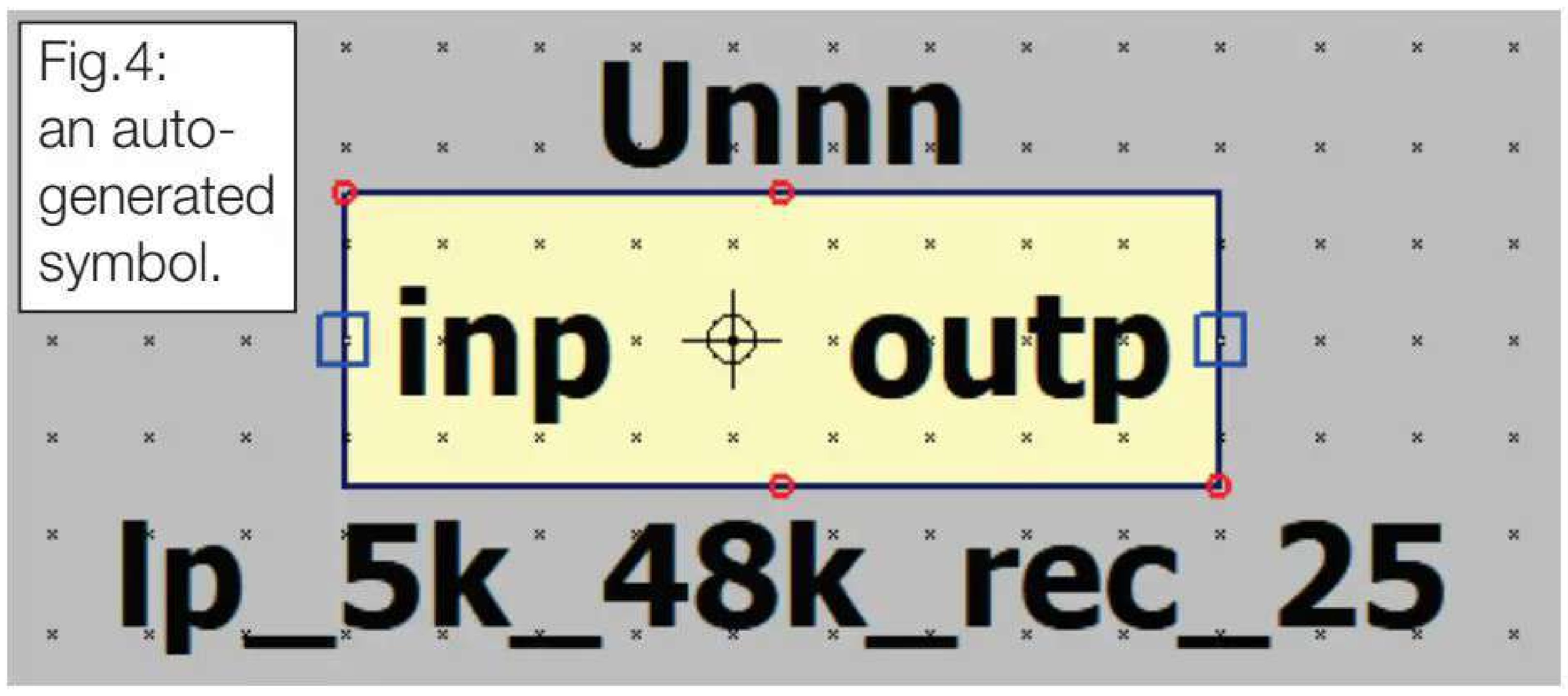
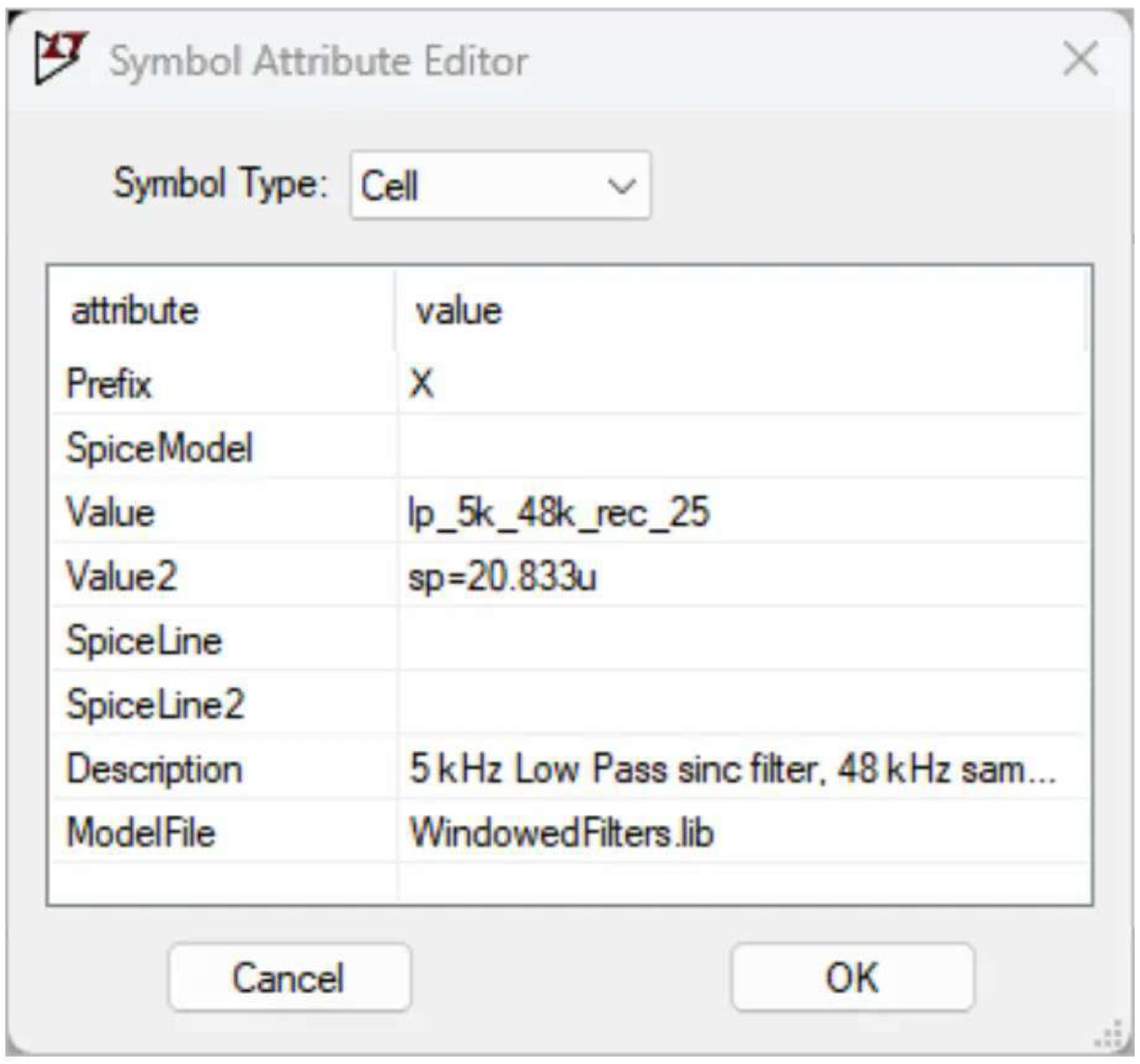


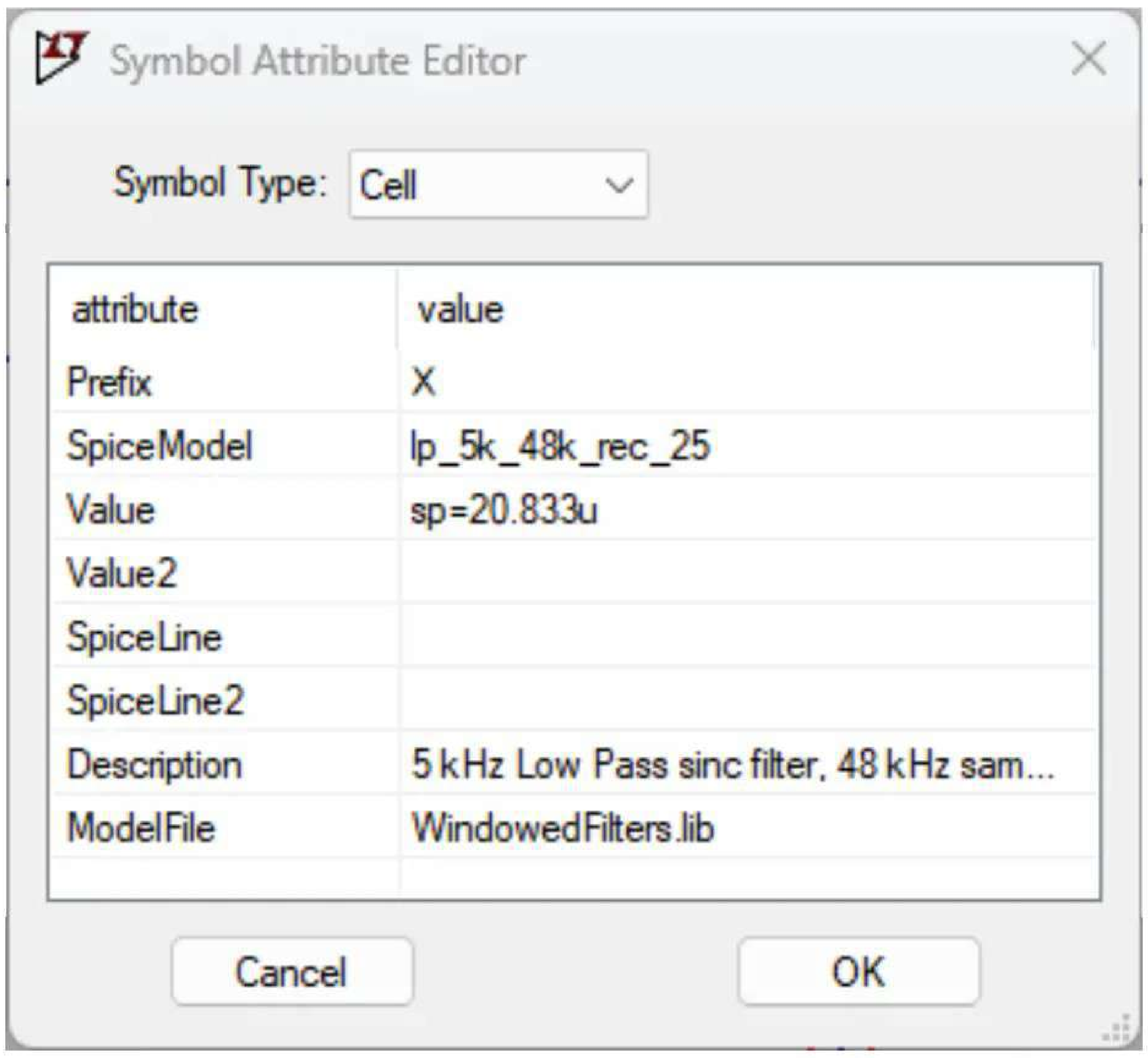
Fig.4: an auto-generated symbol.



Symbol Type: Cell

attribute	value
Prefix	X
SpiceModel	
Value	lp_5k_48k_rec_25
Value2	sp=20.833u
SpiceLine	
SpiceLine2	
Description	5 kHz Low Pass sinc filter, 48 kHz sam...
ModelFile	WindowedFilters.lib

Fig.6: attributes configured without "SpiceModel" defined.



Symbol Type: Cell

attribute	value
Prefix	X
SpiceModel	lp_5k_48k_rec_25
Value	sp=20.833u
Value2	
SpiceLine	
SpiceLine2	
Description	5 kHz Low Pass sinc filter, 48 kHz sam...
ModelFile	WindowedFilters.lib

Fig.7: attributes configured with "SpiceModel" defined.

attributes to make these changes and adding the sample period (**`sp`**) parameter, the attribute editor is as shown in Fig.6. There is also a "Description" attribute; text placed here is displayed in the component tool in the schematic editor.

By default, the "Value2" attribute (**`sp`** in our case) will not be displayed on the schematic when the symbol is placed. It can be made visible by going to the menu item **Edit → Attributes → Attribute Window** in the symbol editor, selecting "Value2" (see Fig.8), then placing the text on the symbol (for example, see Fig.9).

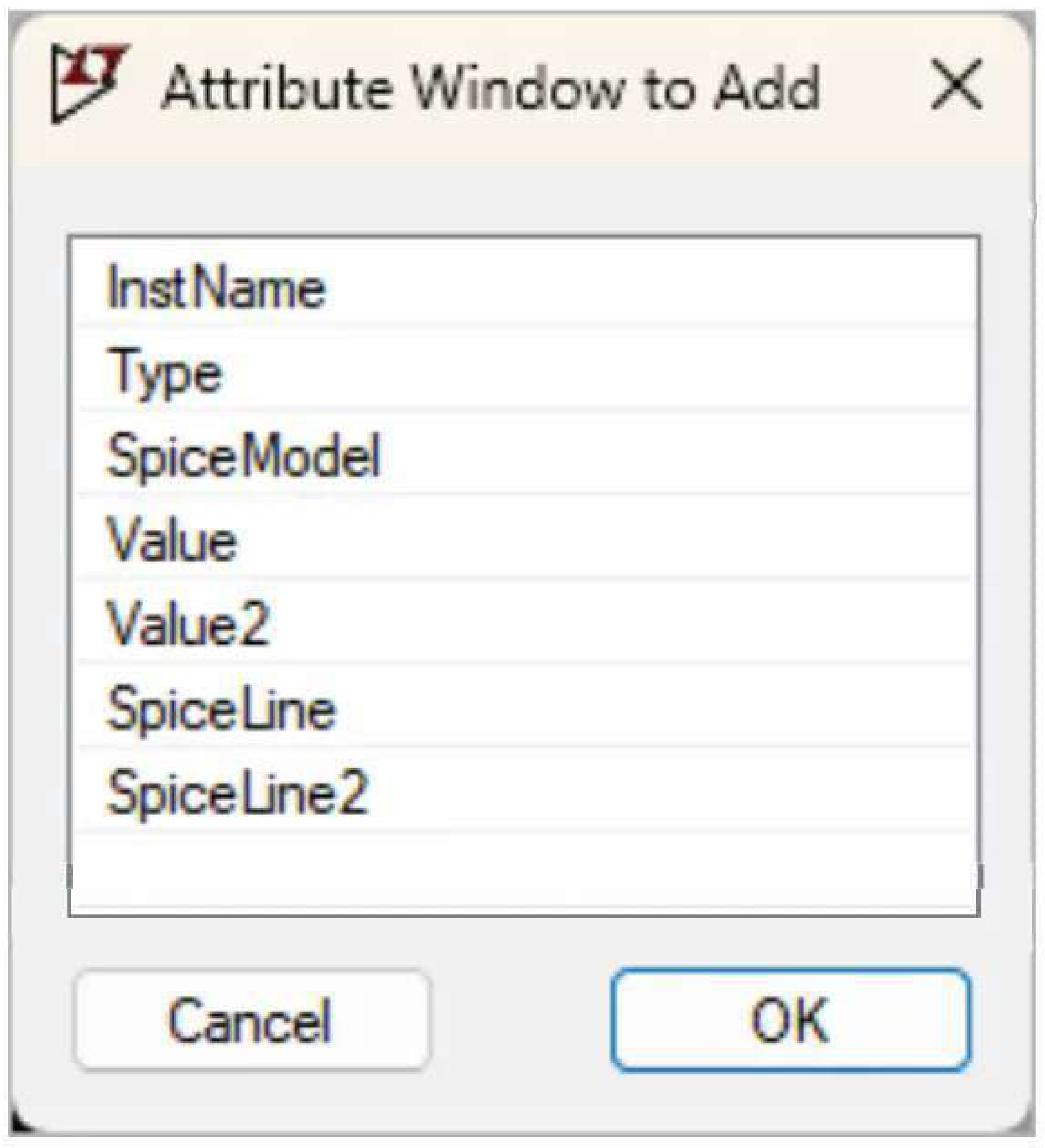


Fig.8: the LTspice Attribute to Add dialog.

The text may be too small when initially placed on a symbol. If this happens, the font size can be changed by right-clicking on the text and selecting a different value in the dialog box that appears.

Setting the subcircuit name to be in the "SpiceModel" attribute, rather than the "Value" attribute (see Fig.7) changes the behaviour of the symbol on the schematic – if there are multiple subcircuits in the "ModelFile", they can selected after the symbol is placed (more on this later). If this is done, the parameters should be put in the "Value" rather than "Value2" attribute.

Copying and editing symbols

An alternative to auto-generating a symbol is to copy one from the library provided by LTspice and edit it. This is a quick way to obtain a symbol with a shape other than a box without having to draw it from scratch. In this case, the lowpass filter symbol for the **`2ndOrdLowpass`** behavioural filter special function (**2ndOrdLowpass.asy**) is suitable.

By default, the LTspice library is in **`C:\users\username\AppData\Local\LTspice\lib\`**

The symbols are in the **`sym`** folder at this location, and the filters are in the **`sym\SpecialFunctions`** folder of the library. Do not edit the symbol in the library; copy it to the folder containing your subcircuit library, change the file name to match the subcircuit name and edit this copy.

To use this symbol for the filters, we need to change pin names to **`inp`** and **`outp`**, change the attributes to match the details of the subcircuit, and edit the information displayed on the schematic by the symbol. For example, the original **2ndOrdLowpass.asy** symbol displays the parameters but not the model name. An example edited symbol is shown in Fig.10.

When editing symbols, schematic files should be closed. LTspice should be restarted after new or updated symbols are ready. If symbols have been changed, they may need to be deleted and placed on the schematic again.

Using your symbols

To place your own symbol on the schematic, use the component tool in the usual way but change the "Top Directory" dropdown to select the folder with the

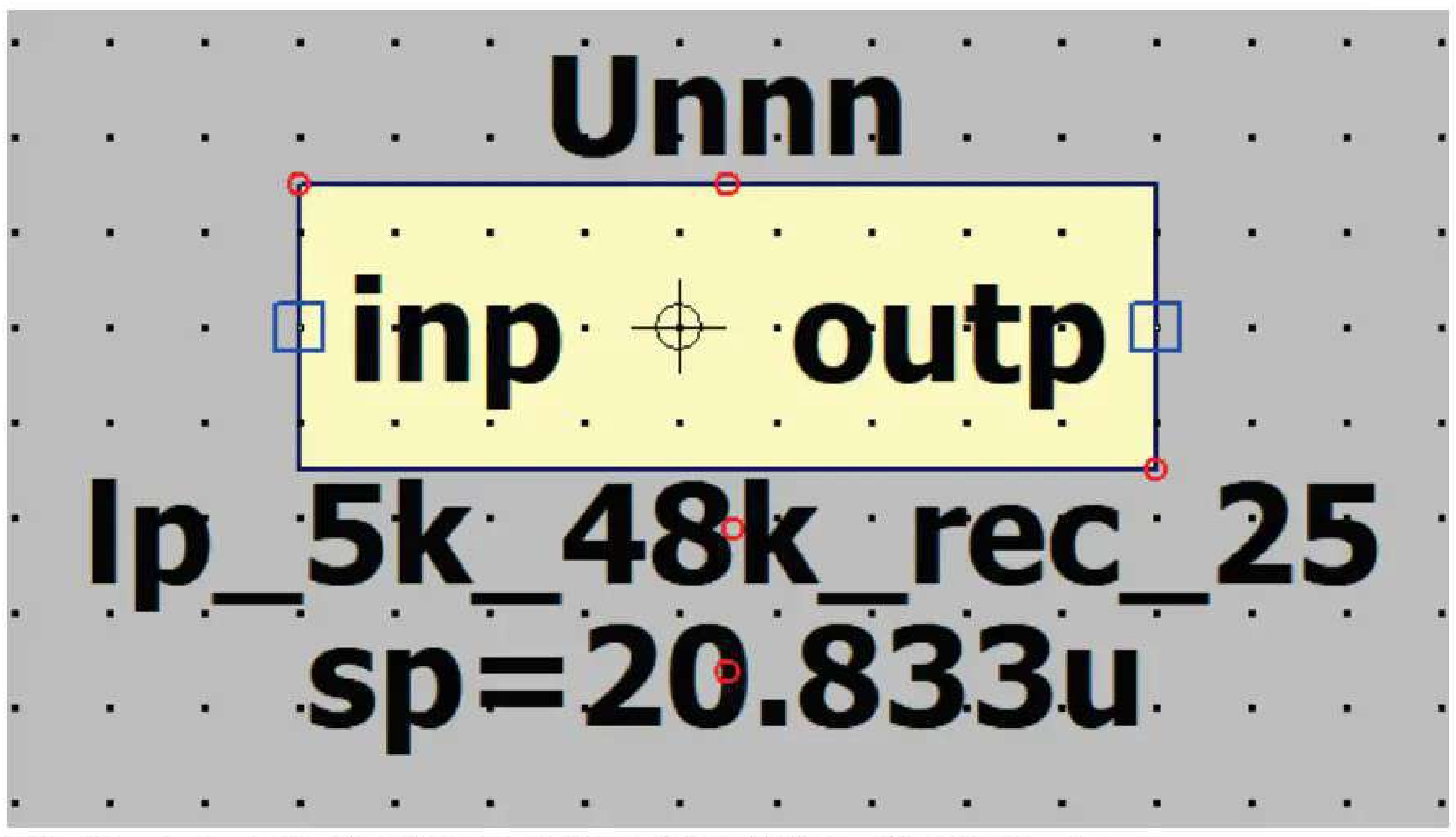


Fig.9: a symbol after adding the "Value" attribute.

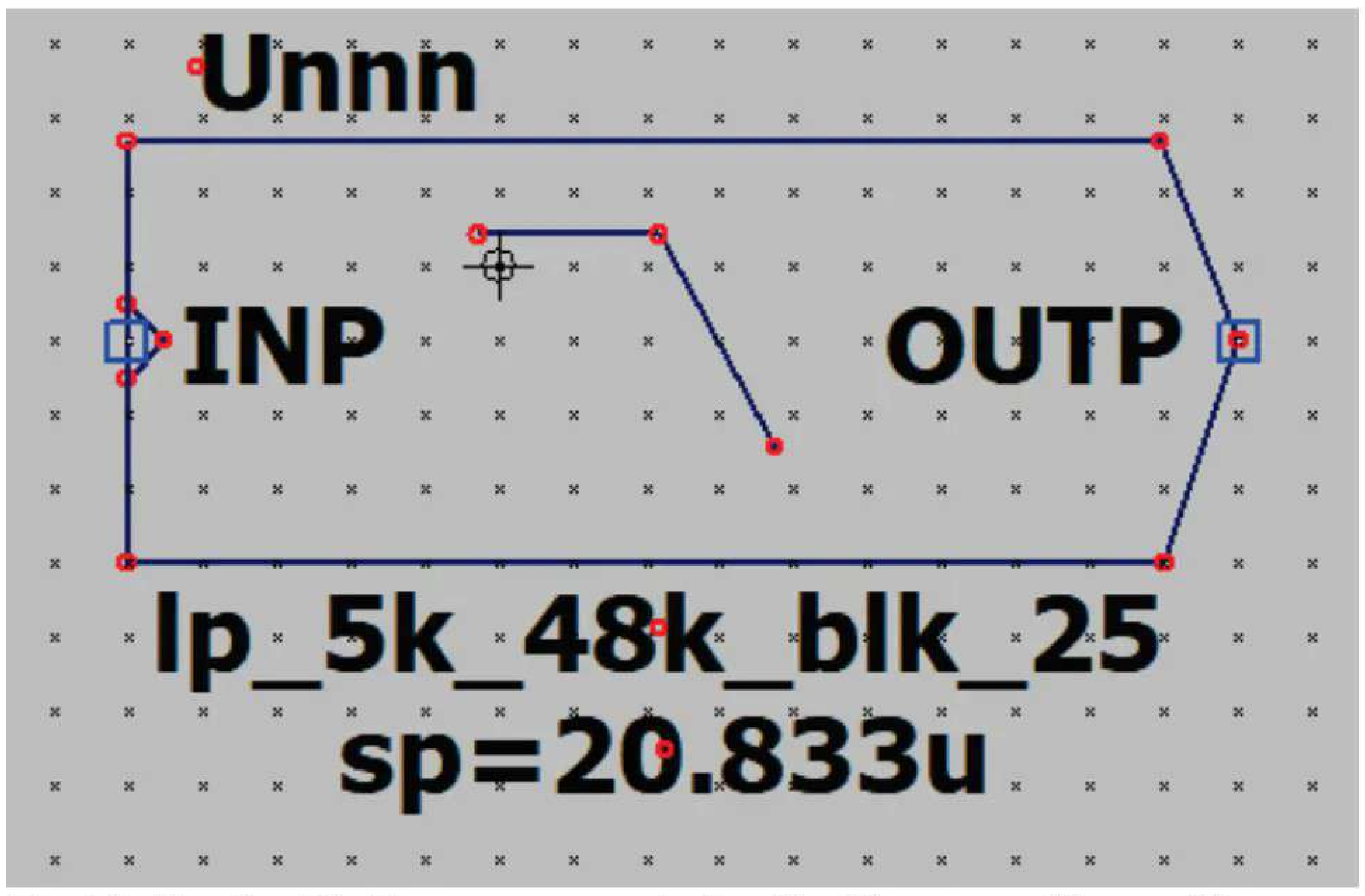


Fig.10: the 2ndOrdLowpass symbol edited for use with our filter.

current schematic in it. If there are symbols there (.asy files), they will be listed (see Fig.11). When selected, the symbol and description of the component should be displayed in the usual way.

If the current folder in the dropdown is not correct, save the schematic, close and restart LTspice, and open the schematic again from LTspice.

By default, the "Top Directory" dropdown provides the standard LTspice library location (in `AppData` as noted above), the current folder and the "user library" folder, which defaults to `C:\Users\username\Documents\LTspice`

The user library location can be changed and additional library locations can be added via the settings dialog at **Tools →Settings → Search paths**. If the symbols and associated library/subcircuit files are moved to the user library, they will be available to add to schematics created in any folder.

Example schematic & simulation

Fig.12 shows a schematic with two digital filters added via symbols. This is similar to the examples discussed last month using text on the schematics. The two filters use the different symbols discussed above to help illustrate these, but for real use, it would make sense to use the same symbol style for a set of related filters.

Note that the `.lib` or `.inc` commands are not required on the schematic as they were for the text-based version, because they are automatically included when the symbols are netlisted.

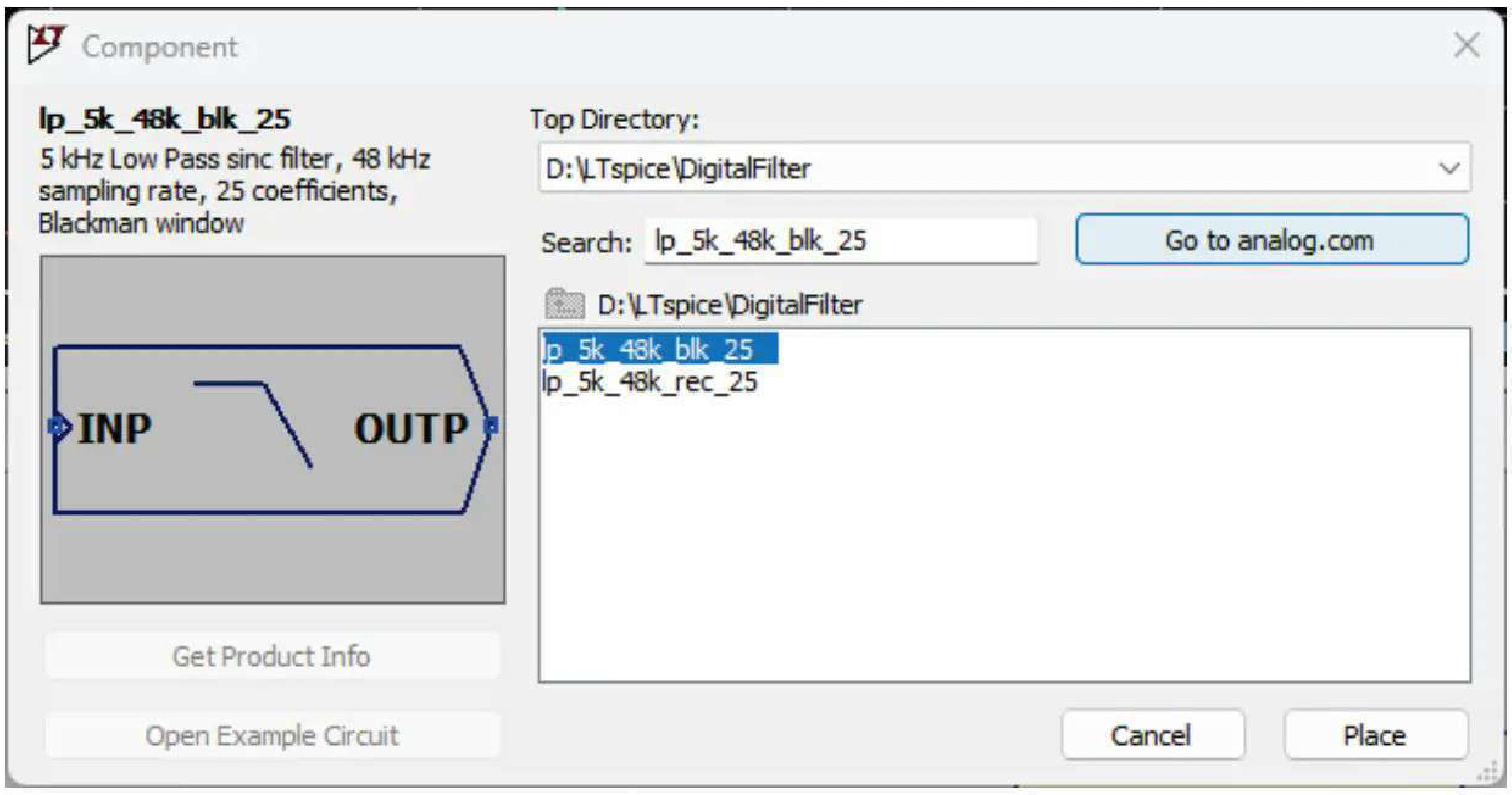
In this example, U1 is the 5kHz low pass, with 25 coefficients, 48kHz sampling and rectangular window sinc filtering, as discussed in previous articles. The `sp` parameter is left as the default value.

U2 is the variant of the same filter using the Blackman window (discussed last month). This filter has its `sp` parameter set to double the designed-for sample period. The filter will still operate but will have half the cutoff-frequency. This was done to illustrate independence of parameter setting.

The simulation results in Fig.13 show the circuit operating as expected, with the cutoff frequency of U2 half of that for U1.

As with standard LTspice components, right-clicking on the symbols of filters allows the values to be edited. In this example, the symbol used for U1 had its attributes configured as shown in Fig.6; U2 used the approach shown in Fig.7. This results in different behaviour from the right-click.

For U1, where the symbol does not have a "SpiceModel" attribute defined, right-click enables parameter values to be edited (see Fig.14). The `sp` parameter can also be edited by right-clicking on the text on the schematic.



lp_5k_48k_blk_25
5 kHz Low Pass sinc filter, 48 kHz sampling rate, 25 coefficients, Blackman window

Top Directory: D:\LTspice\DigitalFilter

Search: lp_5k_48k_blk_25

D:\LTspice\DigitalFilter

lp_5k_48k_blk_25
lp_5k_48k_rec_25

Fig.11: the component selection dialog box.

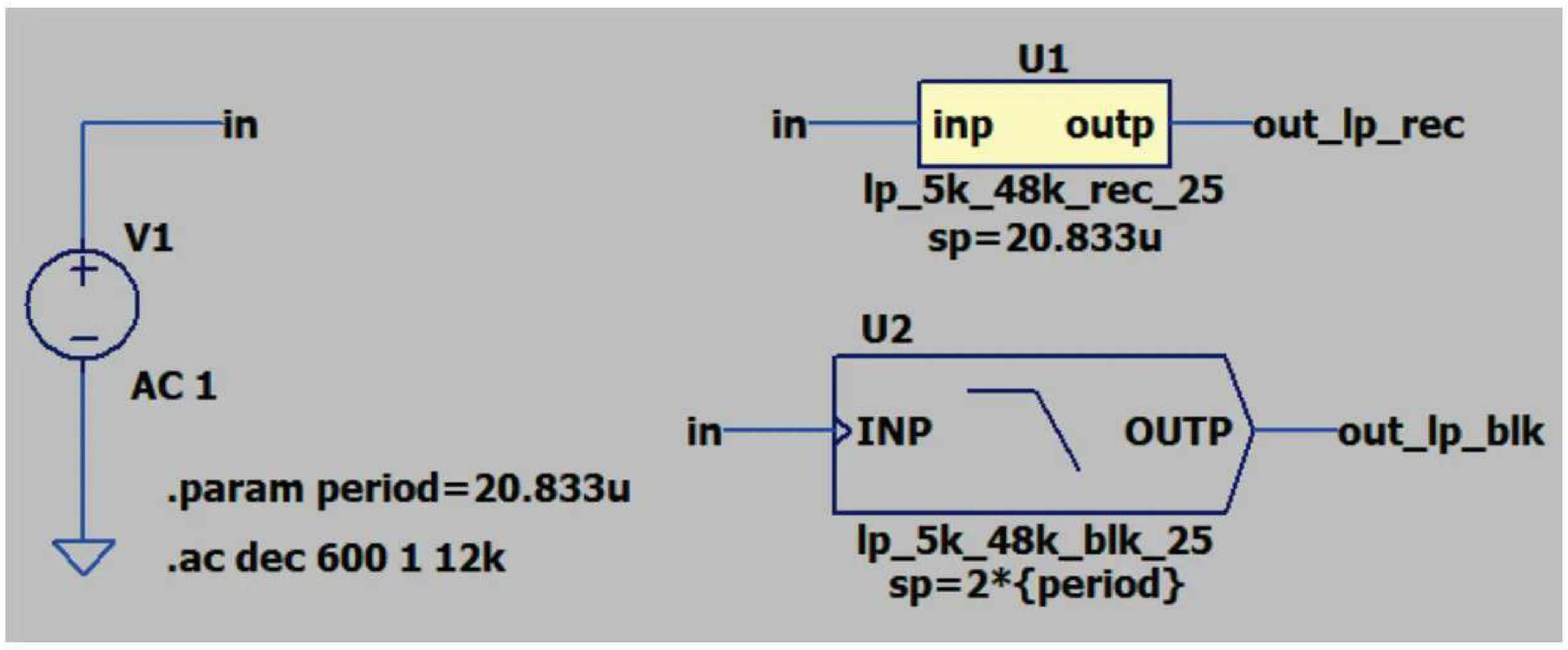


Fig.12: an LTspice schematic using the symbols for filters.

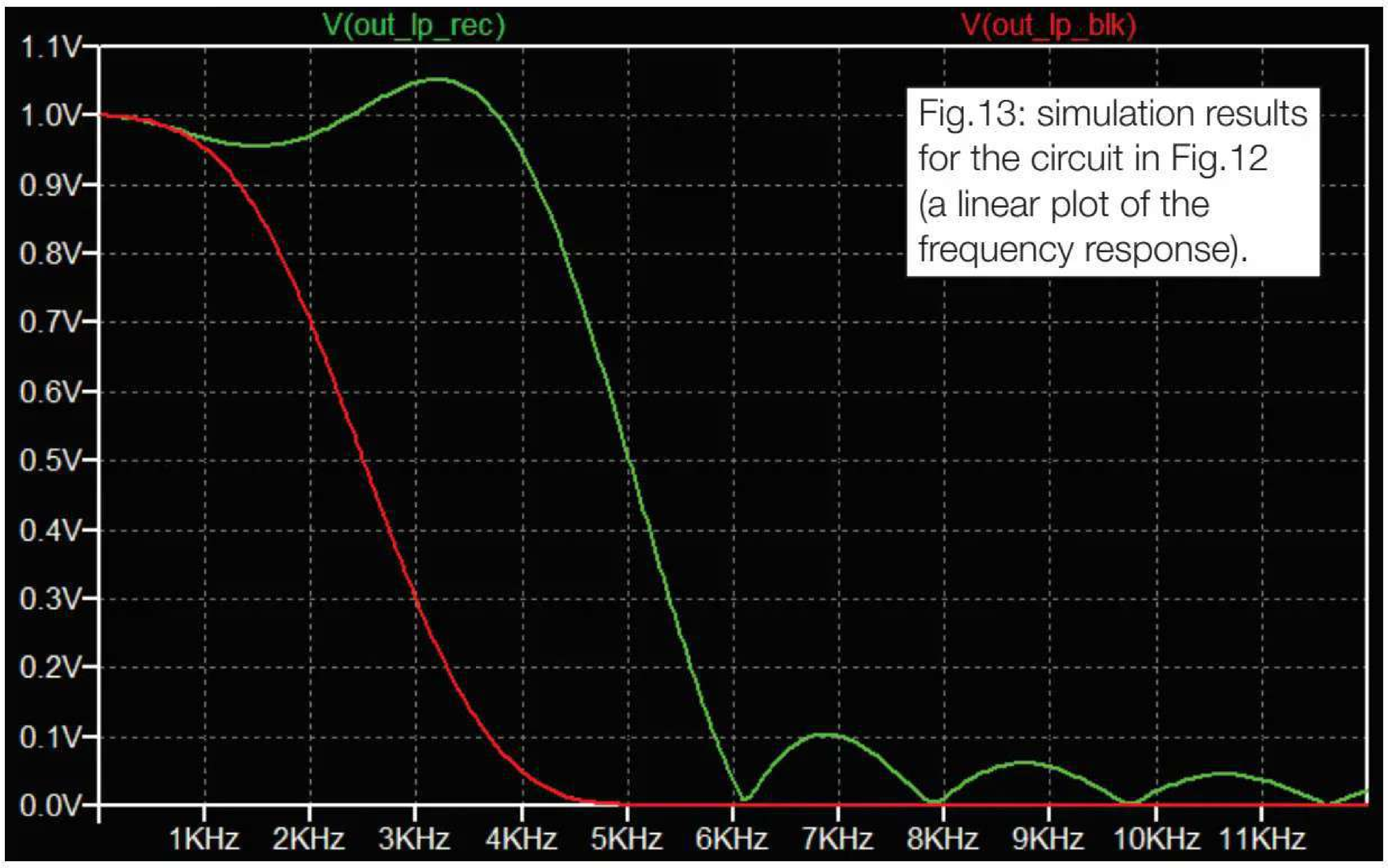
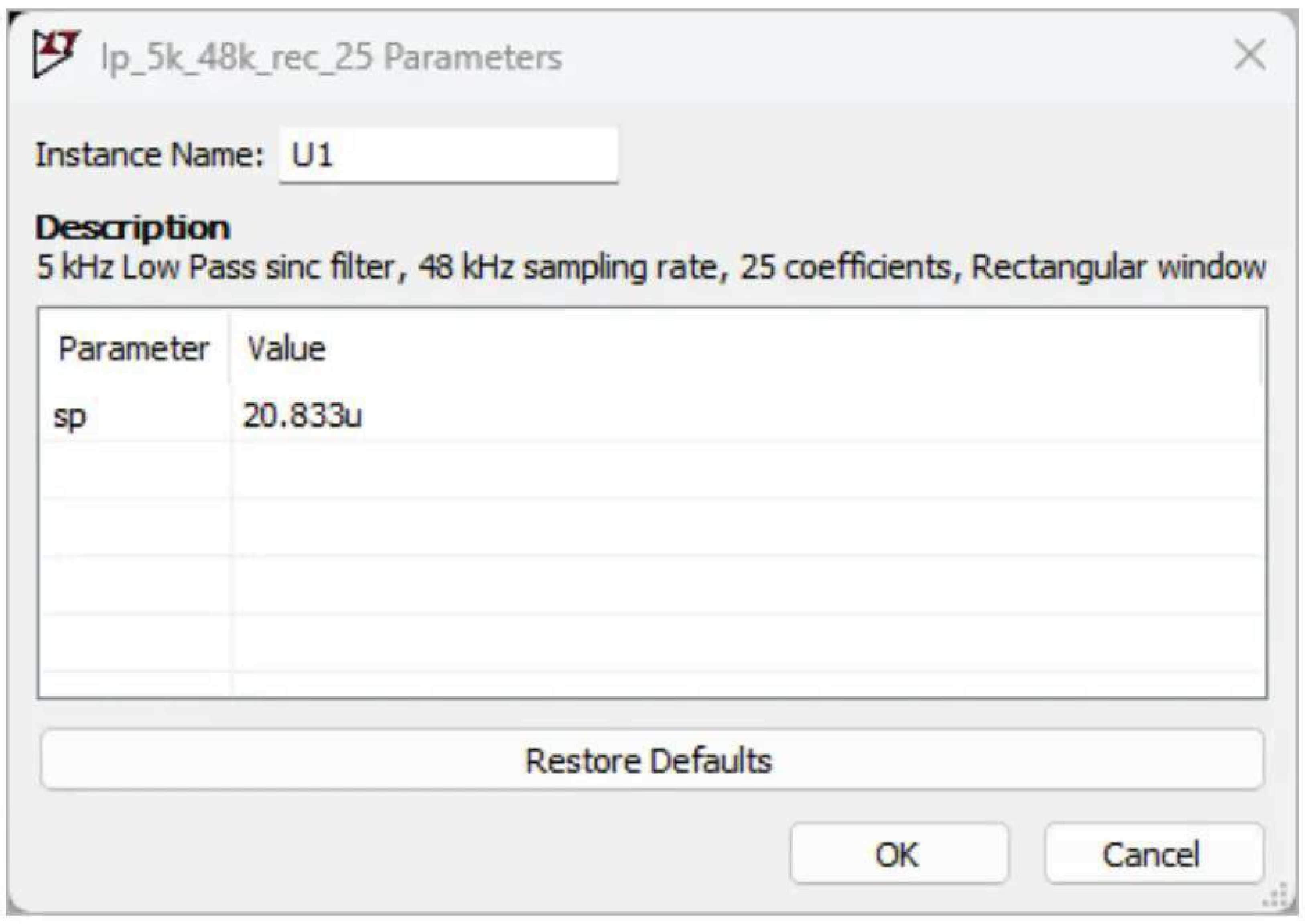


Fig.13: simulation results for the circuit in Fig.12 (a linear plot of the frequency response).

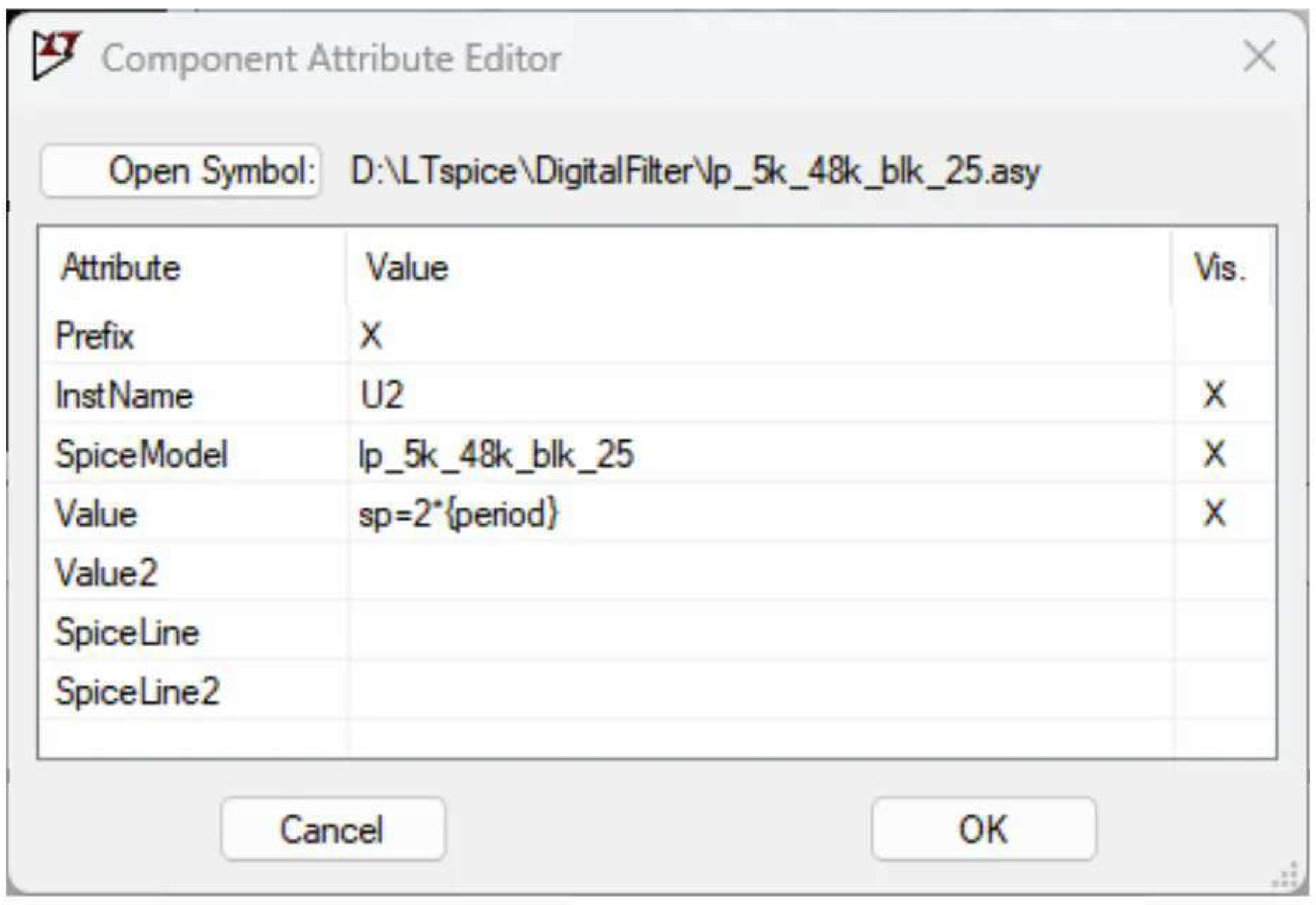


Instance Name: U1

Description
5 kHz Low Pass sinc filter, 48 kHz sampling rate, 25 coefficients, Rectangular window

Parameter	Value
sp	20.833u

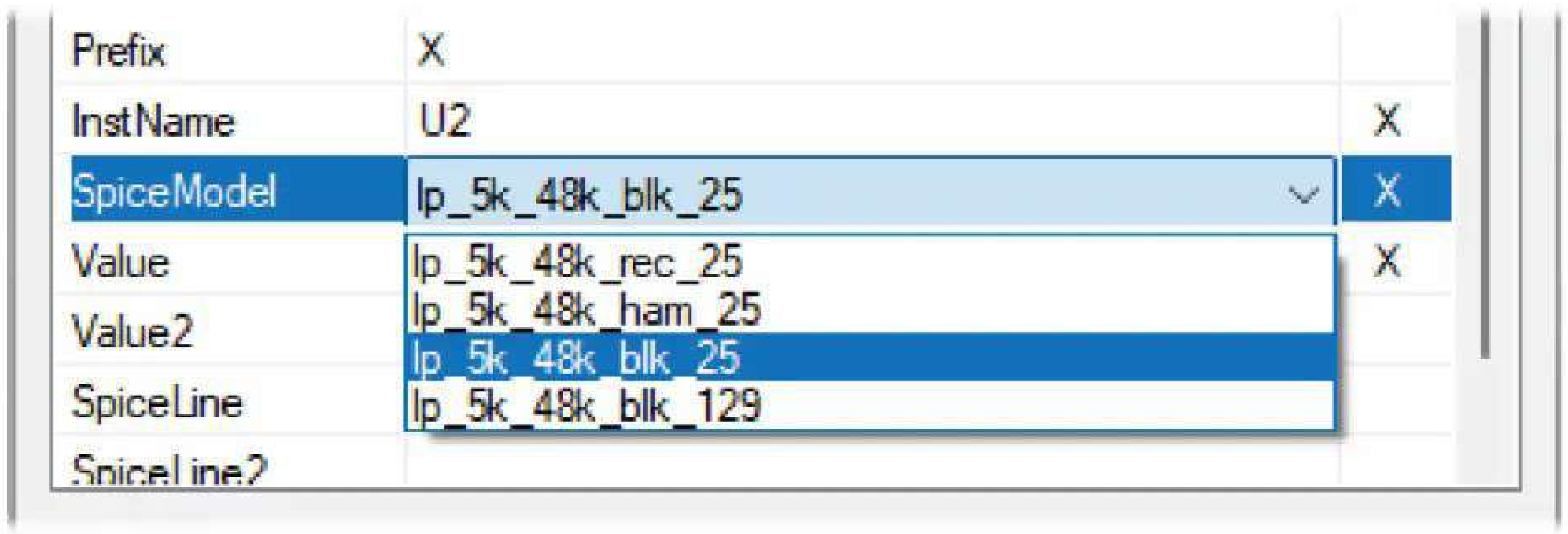
Fig.14: the symbol parameter editor dialog box.



D:\LTspice\DigitalFilter\lp_5k_48k_blk_25.asy

Attribute	Value	Vis.
Prefix	X	
InstName	U2	X
SpiceModel	lp_5k_48k_blk_25	X
Value	sp=2*{period}	X
Value2		
SpiceLine		
SpiceLine2		

Fig.15: another symbol parameter editor dialog box.



Prefix	X	
InstName	U2	X
SpiceModel	lp_5k_48k_blk_25	X
Value		X
Value2		
SpiceLine		
SpiceLine2		

Fig.16: selecting models (subcircuit definitions for a symbol).

For U2, where a "SpiceModel" attribute is defined, right-click enables attribute values to be edited via the dialog (see Fig.15). This approach also allows the model to be changed.

Right-clicking on the "SpiceModel" value in the dialog provides a dropdown list to select from all the subcircuits in the file defined by the "ModelFile" attribute (see Fig.16). This means that only one symbol can be used for a set of related components.

Obviously, all the subcircuits have to be compatible with the symbol, otherwise errors will occur. In this example, the **WindowedFilters.lib** file contains the four filters discussed last month.

With the no "SpiceModel" symbol definition variant (as for U1 here), it is still possible to access the full attribute editor (similar to Fig.15) by holding the control key when right-clicking the symbol. However, this does not include the dropdown list of "SpiceModel" choices – the subcircuit associate with the symbol is fixed.

Impulse response simulations

In our discussions on the windowed sinc filter, we covered the important concept of impulse response. As a brief recap, for a digital filter, the impulse response is the output produced when the input to a circuit is equal to one for a single sample period and otherwise zero for all time. For the windowed sinc filter, the coefficients required to implement the filter are obtained from the impulse response.

The required impulse response can be obtained from the Inverse Fourier Transform of the filter frequency response, and it is well known to be a sinc function in the case of ideal low-pass filters. As this sinc function is non-zero to infinity, the number of coefficients must be reduced for implementation.

Simple truncation of the function (called a rectangular window) is feasible, but better results are obtained by multiplying the sinc function by smoother window functions, like the Hamming and Blackman types. This was discussed in detail over the past couple of months.

Having obtained a working simulation of a windowed sinc filter in LTspice, it is worth asking if we can obtain the impulse response from the simulation. This is useful to help reinforce the concept of impulse response. It will also provide a check that the simulation is correctly set up because we know that the result should match the coefficients used to implement the filter.

There are two ways that we can obtain the impulse response from the simulation. We can apply an Inverse Fourier Transform to the simulated frequency response (such as Fig.13), or we can run a transient simulation with an impulse input. We will look at these in turn.

Fig.17 shows a schematic similar to Fig.12 but with consistent filter symbols, and both filters using their designed sampling rates. These are the 5kHz lowpass filters with 25 coefficients and a 48kHz sampling rate we discussed over the last couple of months (with rectangular and Blackman windows). We will find the impulse response of these two filters.

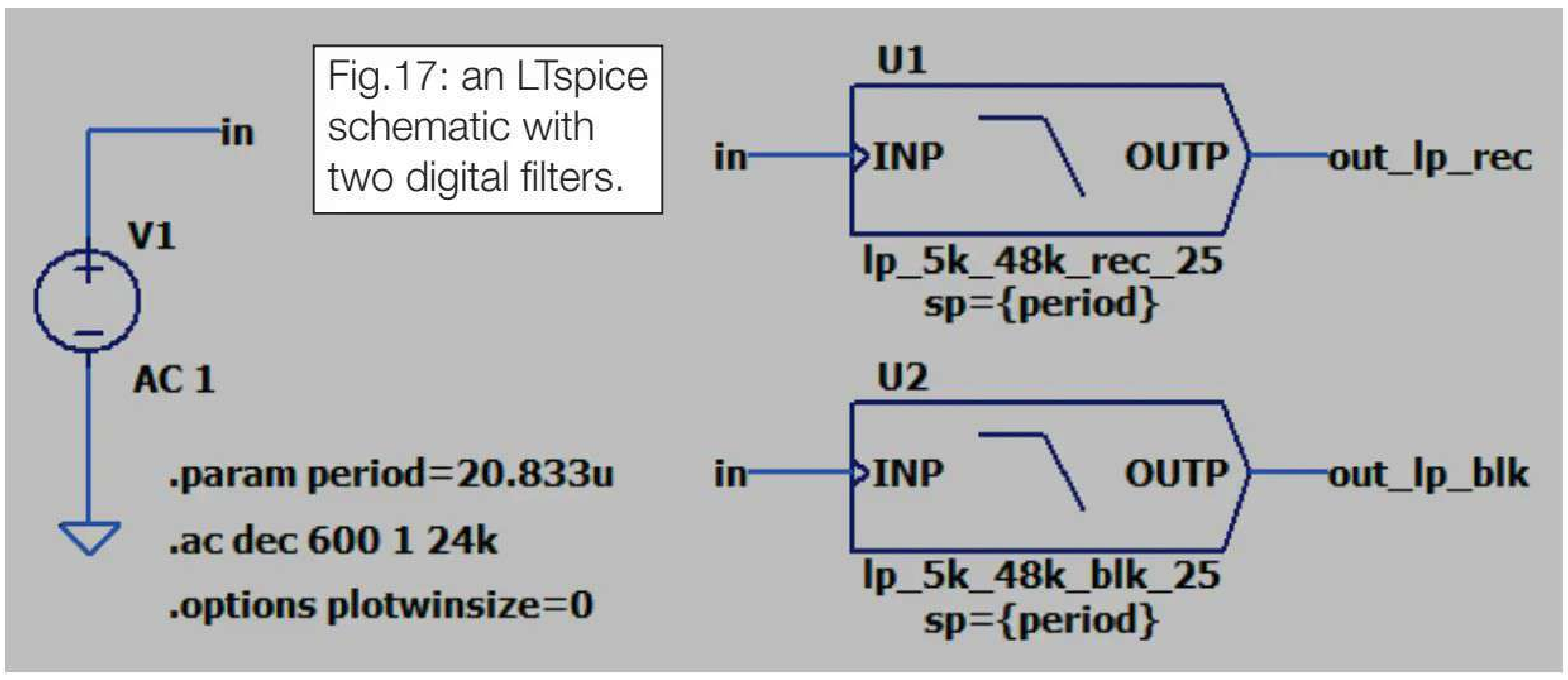


Fig.17: an LTspice schematic with two digital filters.

```
.param period=20.833u
.ac dec 600 1 24k
.options plotwinsize=0
```

Inverse Fourier Transform

LTspice has a Fast Fourier Transform (FFT) that can be applied to the waveforms obtained from transient simulations to obtain a spectrum from the signals. What seems to be less well known is that LTspice can apply the waveform viewer FFT command to an AC simulation result to obtain the impulse function waveform.

The right-click menu is still presented as **View → FFT**, but the dialog that opens to select the signals is titled "Impulse Response", so this is clearly not accidental. However, this inverse FFT (IFFT) does not seem to be documented in the help.

The frequency response results from simulating the circuit shown in Fig.18 should be familiar. Applying the FFT function to this produces results that initially do not look very useful because they extend over a large time range, with a peak of around 10kV at the start.

The numerical results need to be scaled by dividing the sampling frequency (48kHz) to obtain the coefficient values. This can be achieved by right-clicking on the waveform names at the top of the plot and editing the expression to plot to apply the scaling, eg, from `V(out_lp_rec)` to `V(out_lp_rec)/48000`.

The time axis needs to be adjusted to just show the range over which the impulse function is non-zero. As there are 25 coefficients and the sampling period is 20.833µs, this is 24*20.833, which is close to 500µs. Fig.19 shows the impulse function results over this range with a time axis tick setting of 20.833µs to align with multiples of the sampling period.

The data points (which should be the coefficient values) are shown via right-clicking and then selecting **View → Mark Data Points**.

The shapes of the impulse function in Fig.19 look correct, but we can perform a precise comparison by exporting the results data using **File → Export data as text**. This will produce a large file, but we only need the first 25 results. The three columns of data (time and the two waveform values) can be copy-pasted into a spreadsheet.

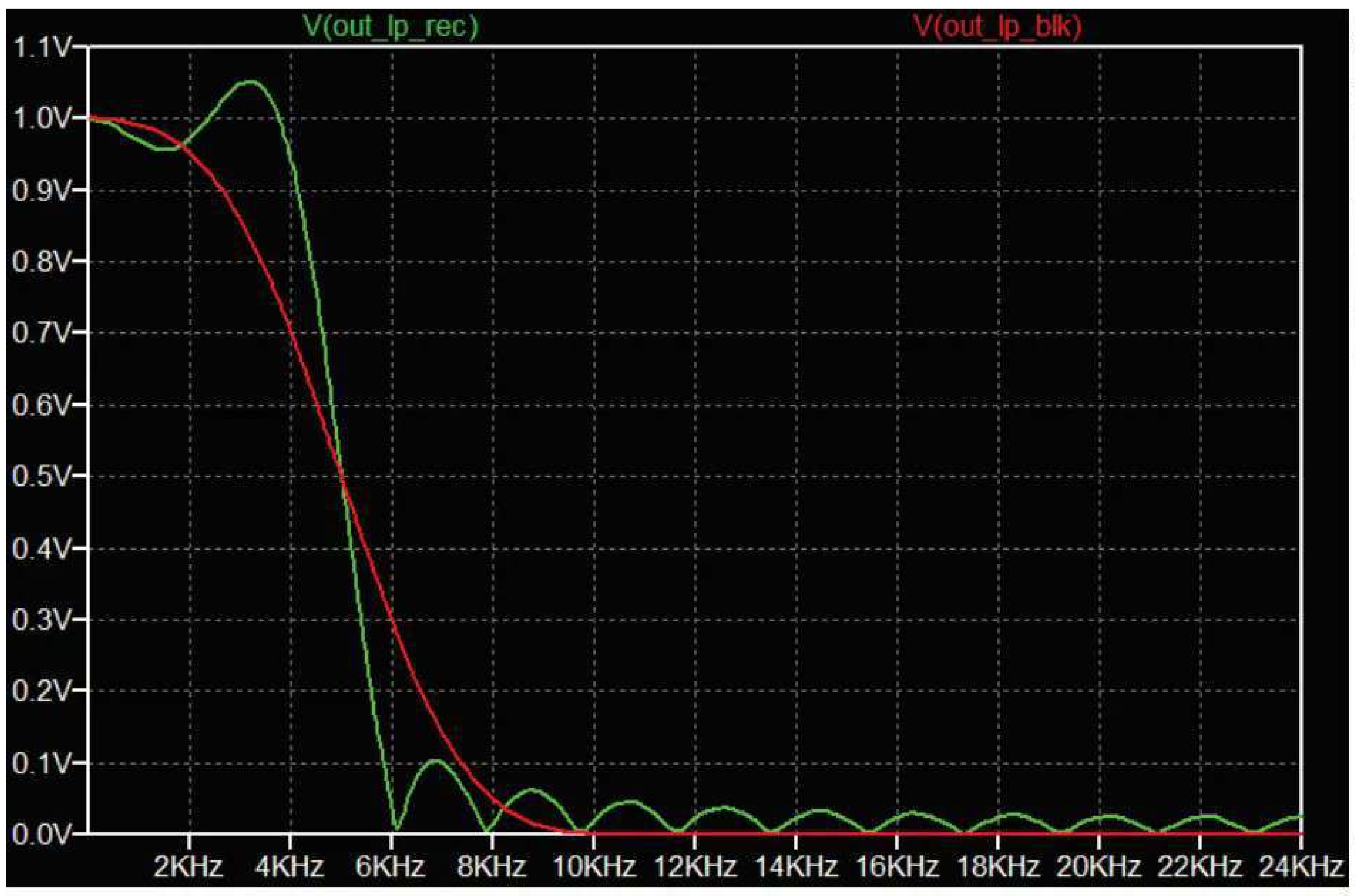


Fig.18: the AC simulation results for the circuit in Fig.17.

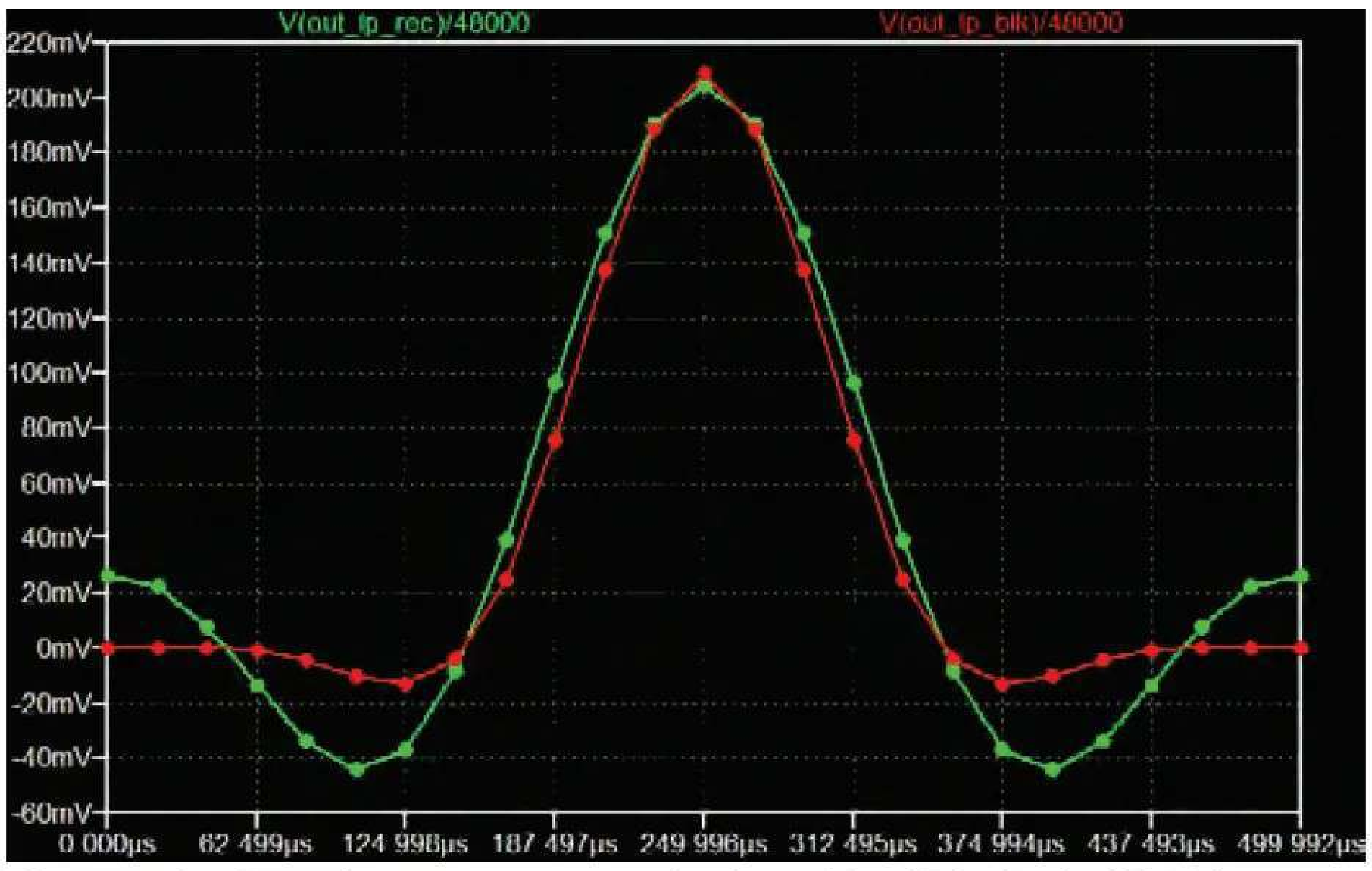


Fig.19: the impulse response calculated by LTspice's IFFT (green trace for the rectangular window, red trace for Blackman).

Fig.20 shows the data for the rectangular windowed filter from Fig.19 plotted with the original coefficient data (see last month) confirming that the IFFT has produced the correct results. The difference from the expected values is below 0.03%.

Transient simulation

We can also obtain an impulse response directly in the time domain using a transient simulation.

We have to consider that our filter models do not behave exactly like a real digital filter in the time domain. This is because their outputs (and internal values) will be calculated at every time step used by LTspice, which will not be aligned to the sample period (the time step in SPICE simulations is adaptive and cannot be set to a fixed value).

This is not a major problem if the input is sampled and held appropriately as we did for examples in *Circuit Surgery*, December 2024 & January 2025, which used a behavioural sample-and-hold component.

For an impulse response simulation, we do not have to add a sampler because we just need one input pulse of 1V amplitude and a duration equal to the sample period. This is straightforward to achieve with a voltage source using a **`PULSE`** or **`PWL`** waveform configuration.

Fig.21 shows the input signal source (**`V1`**) and analysis request from Fig.17 reconfigured for a transient simulation of the impulse response. The pulse waveform is configured to have an amplitude of 1V and a duration of 20.833μs, with a single occurrence (**`Ncycles = 1`**).

This pulse occurs one sampling period after the simulation start, rather than straight away, as it is clearer to see the pulse on the plot and helps the simulation initialise with everything at zero.

The results are shown in Fig.22. The output waveform has the stepped nature expected of a sampled-and-held signal. As this is representing a digital filter, each step on the waveform corresponds to a single numerical value in DSP hardware or software.

LTspice simulations with abrupt steps can produce glitch artefacts (eg, the spike at 41.666μs) that do not relate to the digital filter operation. This problem is reduced by using a small maximum time step (1ns in this case) at the expense of a slower simulation.

The output waveforms appear to have the correct shape and values. A check of one of the numerical values (using the waveform measurement cursors) confirmed a match to the coefficient to the eight digits given by the measurement. This is not unexpected, as the simulation uses idealised behavioural sources.

The IFFT involves a significant amount of processing of data that was from a previous simulation, so it is not surprising that it is slightly less accurate. **PE**

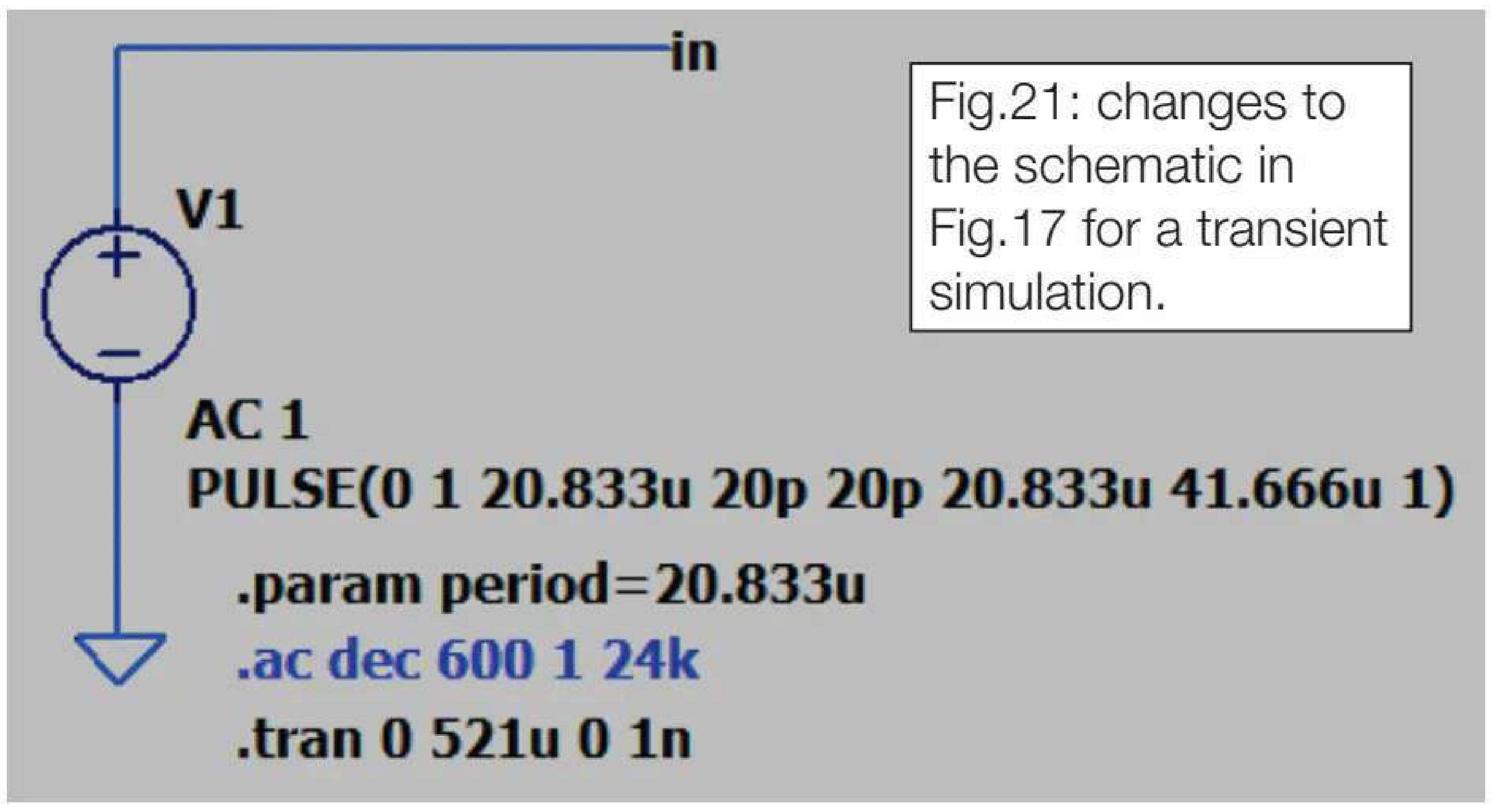


Fig.21: changes to the schematic in Fig.17 for a transient simulation.

```
AC 1
PULSE(0 1 20.833u 20p 20p 20.833u 41.666u 1)
.param period=20.833u
.ac dec 600 1 24k
.tran 0 521u 0 1n
```

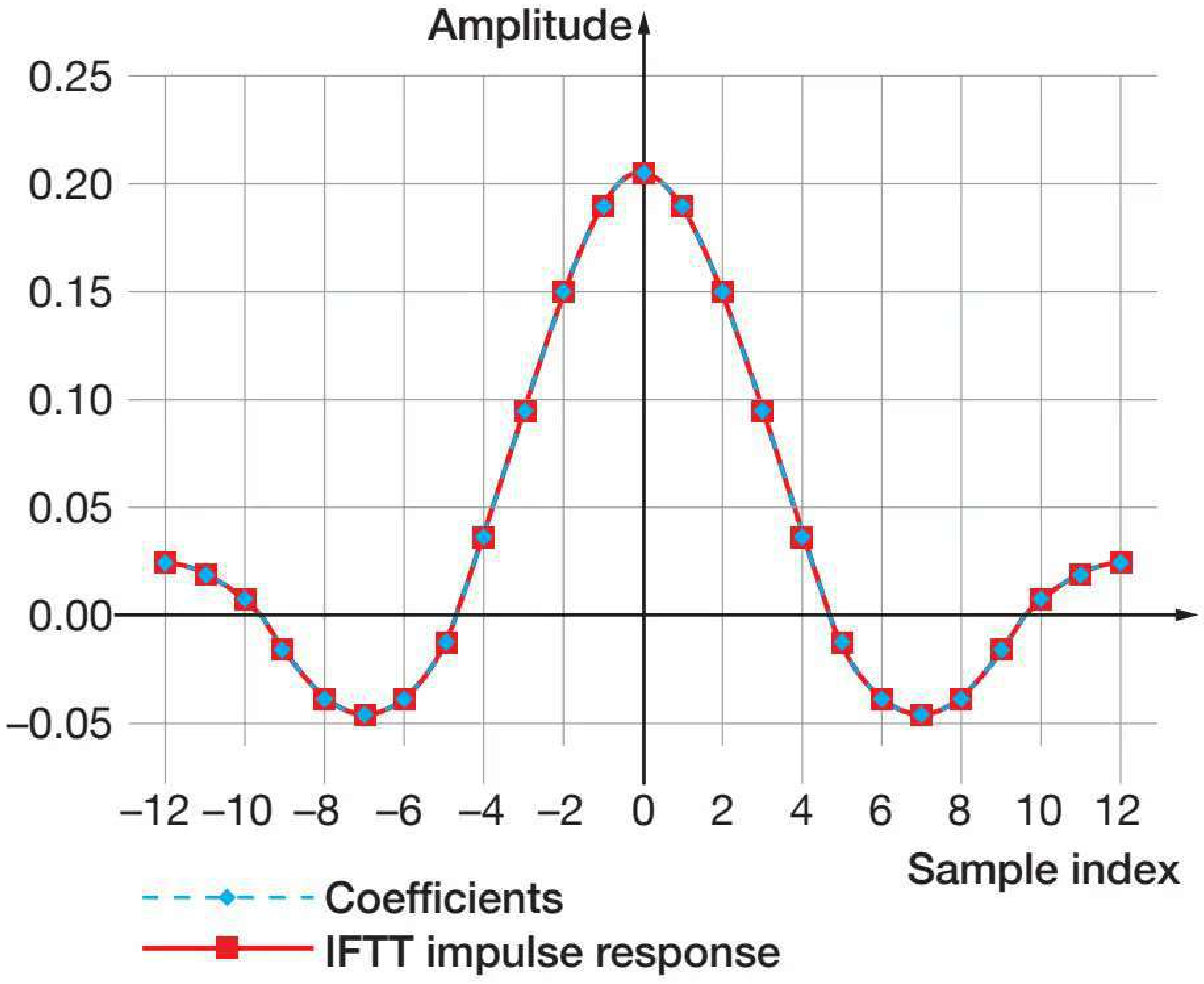


Fig.20: LTspice IFFT results for the rectangular windowed filter.

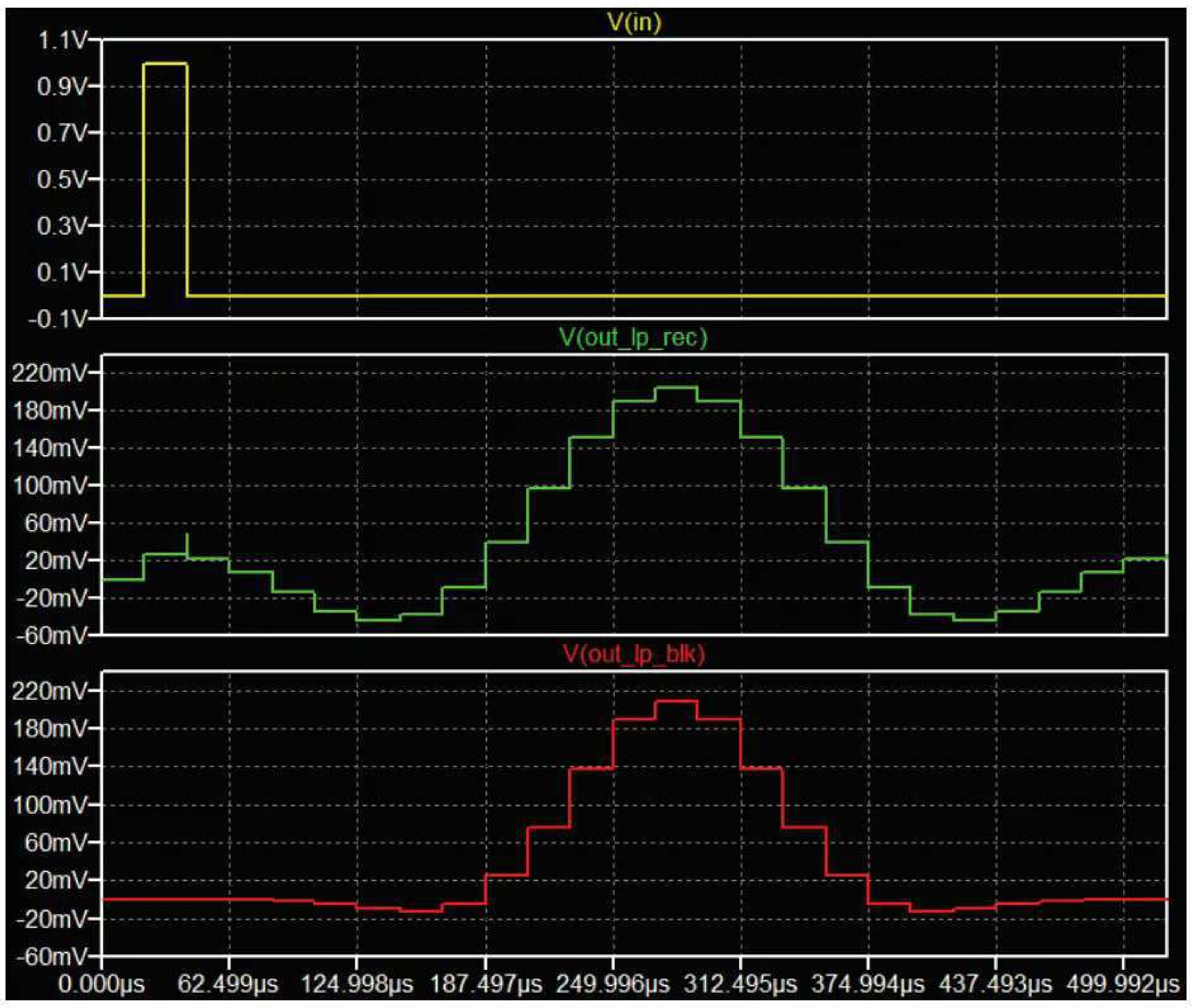


Fig.22: results from the transient simulation of impulse response.

The Fox Report

Barry Fox's technology column

Trying Linux Mint

We are now into the year when Microsoft is cutting Windows 10 users adrift by ending official security updates – unless the PC owner pays heavily for special treatment.

It remains to be seen whether the grey economy finds a way to share these purchased updates between different users.

Whatever happens, awareness of the problem is spreading and resentment is mounting. Best estimates are that some 250,000,000 Windows 10 PCs will be made obsolete, which is an awful lot of people looking for a fix.

As previously recounted, I have had success (so far) with forcing an update by editing the Registry of a Windows 10 PC that meets most of Microsoft's security requirements, but for reasons that are unclear, fails the CPU test.

Microsoft is now warning that in cases like mine, Windows will refuse to provide updates. So far, my forced PC has been successfully updating. I suspect that this is because Microsoft will tolerate a slightly underpowered CPU, but not more security-centric inadequacies.

It could also be because Microsoft now realises that a lot of users won't just obediently buy a new Windows PC; they will look for alternative operating systems that will run on their existing PCs. Certainly, I have been researching alternative ways of keeping my hardware going if I have to cut it loose from Windows.

The obvious prime candidate is to ditch Windows altogether and install Linux, as several of my computer enthusiast friends keep telling me to do. And Microsoft's hostile Windows 11 upgrade policy could be the jumpstart needed for widespread switchover to Linux.

I've previously tried Linux, in the shape of Ubuntu, but for an easy, lazy life have always returned to Windows. So, posing as a non-tekkie Windows 10 user, blocked from upgrading to Windows 11, I tried again.

First, I needed to know which of the many available versions and flavours of Linux to try. I have not yet found any website or posting that directly advises on which version and flavour of Linux is the closest to Windows out-of-the-box.

After some research, I chose the most recent Cinnamon version of Linux Mint (https://www.linuxmint.com/download.php). Perhaps there are better choices; I can't know without trying them all, which I cannot do.

The Mint download is an ISO file that must be written to a bootable USB flash drive (or DVD). The Mint site tries hard to help newbies through the need to download and install the correct Etcher "asset" to do this.

Other sites suggest installing different software, such as Rufus. I tried and found that the required site is (as often the case) cluttered with big green button "Start Now" click traps that suck innocents into downloading completely different software.

Windows warns the user about the risk of 'harming' their computer if they continue with downloading and installing software not approved by Microsoft. At a time when we are all being told to be on our guard against inadvertently downloading malware, it takes experience and skill or brave stupidity to click past these warnings.

Writing to the USB with Etcher is made as easy as possible, but the user has to find where their PC has stored the ISO file (which can vary) and then click to write it to the correct target, not mistakenly overwrite another attached drive.

Armed with the bootable USB, the user has now to know how to temporarily change the boot sequence at BIOS level to force the PC to boot from the USB rather than Windows (unless that is the default). The choice of what function keys to tap at the right time and speed during Windows PC startup varies with the brand.

In the case of my new Acer, repeated attempts at forcing USB boot by tapping the designated boot key (F12) failed with the message "Keyboard Error". Pressing Del for BIOS setup also failed. At first, I thought this might be because I am using a wireless keyboard and mouse (Logitech), because plugging a wired keyboard directly into the PC let F12 open the Boot Menu.

But I cannot simultaneously plug in a USB flash drive because there is only one full size USB socket on the front of the PC. Using a passive USB doubler for a wired keyboard and flash drive still gives the Keyboard Error. I can't hot-swap them either.

Replacing the wireless keyboard dongle with a wired keyboard gave the same Keyboard Error. So I tried with another older PC, which has a wired keyboard and mouse. This booted into Mint with no problems.

I now think it's likely that my use of a KVM switch (which lets me share keyboard, mouse and video monitor between two PCs) blocks F12 access to the Boot Menu (and BIOS).

None of these problems is insoluble, but they will make a lot of Jack and Jill users give up.

When it all works, and the PC boots from the Linux stick, the Mint screen shows three square icons on a dark background. One icon is to Install Linux Mint. I suspect this means it will install to the hard drive, wiping Windows. If so, it's a big pitfall for the uninformed that needs a big red flag.

The other two icons are Computer and Home. Clicking Computer shows square icons for assets like attached drives. The screen looks very different from Windows.

Clicking Home shows green square icons for Desktop, Downloads, Music etc. These do look a bit like Windows. Clicking available options by trial and error may perhaps kick Mint into some kind of Windows-like shape. I haven't yet mustered the enthusiasm to try.

Like assembling flat-packed furniture, it would all be easier next time. But, like assembling flat packs, most people will only be needing to do the job once. And on their learning curve they may end up over-writing Windows by "installing" Mint.

Personally, I believe the big breakthrough for Linux will be when someone offers a version that defaults out-of-the-box to a Windows-like screen and has clear warnings of pitfalls like inadvertently wiping Windows. Time will tell whether anyone does this.

Now there is a new dark horse lurking in the wings, if you will pardon the mixed metaphor. The word from the herd is that Google is working towards a version of the Android phone operating system that can be used on desktops and laptops. This would effectively replace the Chromebook OS, which has hardly been a roaring success for Google.

If 'Android for PC' were packaged as an easy-to-install replacement for Windows, the options for waving goodbye to Windows would change overnight.

The **Fox** *Report*

Barry Fox's technology column

At last, a real torch!

To misquote Crocodile Dundee, "That's not a torch. **This** is a torch!"

Over the years, I have bought many torches (or "flashlights") and have thrown most of them away. The switches fail. Connections corrode. They mysteriously only light up after just the right kind of shaking. And they eat expensive batteies.

So I jumped at the chance to try one from Sandberg, the Danish company that has carved a nice niche for making satisfyingly solid bits of electronics and IT kit, and has now branched out into torches.

Apparently, it has been a year and a half since Sandberg introduced its first torch. Rather scarily the company attributes its "huge success" to "an increased focus on increased awareness of the need for prepping". If and when there is a catastrophic disaster or emergency, or civil war, or just a grid blackout, the first thing everyone needs to be have on hand is a reliable light source.

You also need a guaranteed light source when hiking in the wilds or navigating the streets in a city that is saving money by switching off the street lighting. It's also a good idea to have one in case your car breaks down on a country road at night.

Sandberg is now expanding its 'prep' range of super-rugged torches to four models, all weatherproof with a five-year warranty. Three of them can be used as power banks for charging a flat mobile phone or other USB-powered accessory.

Erling Hoff Petersen, Sales Director at Sandberg A/S, says his torch "lights up an entire horse paddock!".

I don't have a horse or a paddock, but I had a chance to try Sandberg's Active Survivor 7-in-1 Torch Powerbank 10000 (Item no: 421-09) on a dead black night at a deserted sea beach. I can vouch for the fact that it's more like a searchlight than a torch! The front lens telescopes to tighten or widen the beam.

Pressing the main button once gives full brightness; press again for a bit less light, and again for a flashing an "I'm here" signal. Pressing the same main switch more rapidly cycles a small flat LED between bright, not-so-bright and emergency red steady and red flashing.

Be warned, though, that there is a knack to pressing at just the right speed to toggle between the main beam and flat panel modes. At first, I thought the switch was faulty, but I persisted on the assumption that the world would probably end before Sandberg sells a device with a dud switch.

The body is tough aluminium, and the lens is shielded with large teeth that can be used as a hammer to break glass in an emergency, for instance, a sinking car. This also has the benefit that if you put the torch bright-end-down on a flat surface, it will illuminate the general area with the light leaking out between the teeth.

Quality does not come cheap. Prices range from £16 to £54 (including VAT) for the 10Ah/1000 lumens model. All models use lithium-ion rechargeable batteries.

For the same price, you could buy half a dozen Mickey Mouse torches that are guaranteed to let you down when you most need them. PE

Sandberg
ACTIVE
5
Survivor
7in1 Torch
Powerbank 10000
• Powerbank for your smartphone
• Adjustable light intensity
• Zoom lens
• Powerful lantern and flashlight
10000
Double Quality Tested

UP
<S1
DOWN
<S2
MODE
<S3
HOLD FOR SETTINGS
SILICON CHIP
ESR TWEEZERS
©2024
04105242
Tweezers Panel
ESR 3.2V
0.02Ω
S1
S2
S3

TIM BLYTHMAN'S

ESR TEST TWEEZERS

We have produced a few variants of our Test Tweezers since the original version in the October 2022 issue. Still, none has yet had the handy feature of measuring capacitor ESR (equivalent series resistance). Our new ESR Test Tweezers can measure ESR and capacitance while being significantly more compact than all our previous ESR meters!

ESR (equivalent series resistance) is an inherent but undesirable property of capacitors that acts like a resistance in series with the capacitive element. Fig.1 shows this and the other factors that can be used to model a real capacitor, as opposed to an ideal, purely capacitive one.

For good performance, especially at high currents (as in a switch-mode supply), a capacitor's ESR and ESL (equivalent series inductance) should be low and the leakage resistance should be high. That combination best approximates an ideal capacitor.

Generally, the ESL is relatively small and is often lumped together with ESR by specifying it at a known frequency, often 100-120Hz or 100kHz (the former being relevant when rectifying mains AC). The total series impedance can then be specified in ohms.

If the ESR is high, the capacitor will dissipate a significant proportion of the energy that passes through it, unlike purely reactive elements such as ideal capacitors and inductors, which have no losses. It is well known that high-ESR electrolytic capacitors can cause problems, but they are not the only type of capacitor that can suffer from high ESR. Other types, such as plastic film, can be affected too.

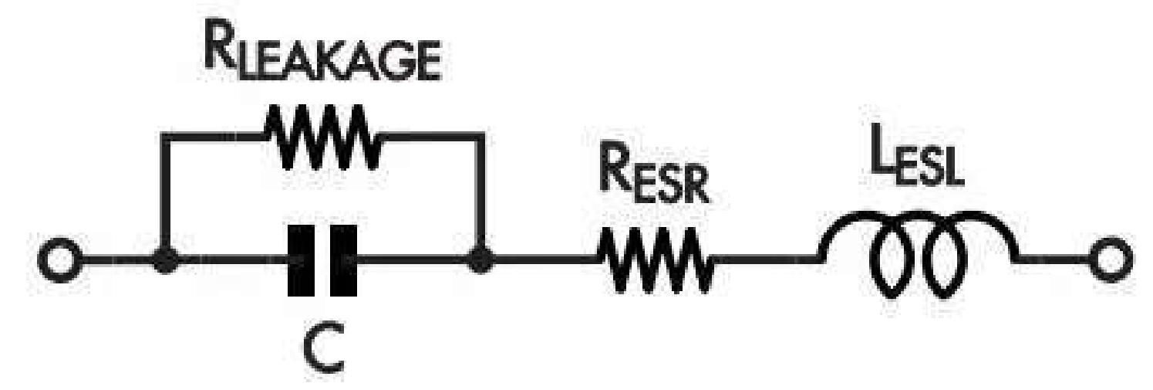


Fig.1: the behaviour of real capacitors, especially electrolytic types, deviates from the ideal model of capacitors found in textbooks. ESR (equivalent series resistance) is one of the more prominent unwanted phenomena; capacitors often fail due to the ESR rising to unacceptable levels.

In a power supply, a high ESR manifests as a voltage drop due to the current flowing in and out of the capacitor. That will decrease the voltage available to the circuit and heat up the capacitor, sometimes to the point that its contents boil and spill out!

Electrolytic capacitors depend on an electrolyte as the current path between the oxide dielectric layer and the cathode. If this electrolyte dries out, its resistance and thus the ESR will increase. Increasing ESR will also cause an increase in dissipation inside the capacitor, further heating and drying out the electrolyte.

A high ESR capacitor will often cause mysterious or intermittent faults, as documented extensively in our Serviceman's Log pages, where replacing the electrolytic capacitors usually fixes a power supply. The conductivity of the electrolyte can also change with temperature, leading to problems that appear or disappear as the capacitor heats up after the equipment is turned on.

In audio circuits, the higher-than-expected ESR can change the frequency response of a circuit and may increase distortion. These are just some of the scenarios where a high ESR can cause problems.

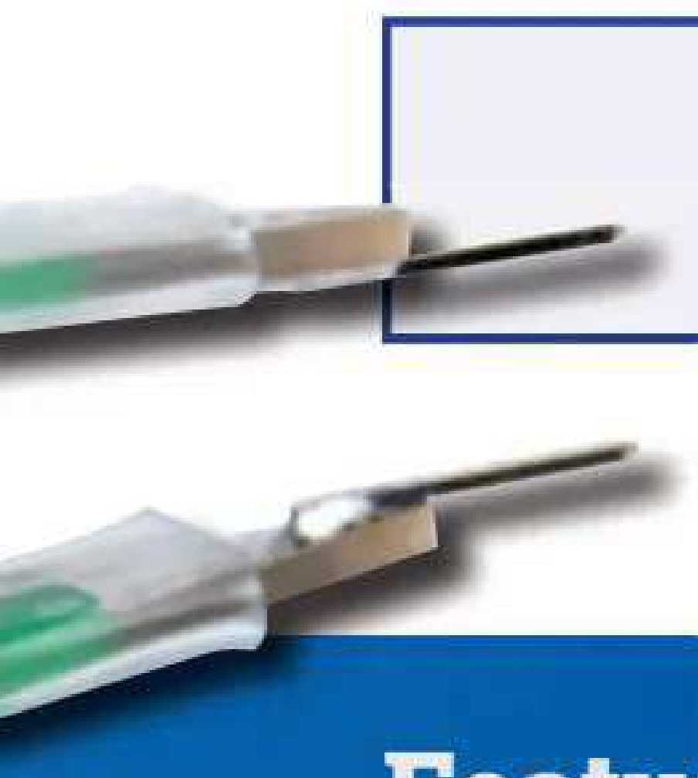
If you have a device that has failed or isn't working correctly, after checking for obvious visual faults like burned components or failed solder joints, the next step is usually to test the electrolytic capacitors. If any are found to have a low capacitance, high leakage or high ESR, they may well be the culprits. Often, several are found to be on the way out.

So, an ESR meter is a very valuable piece of equipment for making repairs and even checking new components to verify that they will perform as expected.

One frequently-seen piece of advice is a warning not to connect the ESR Meter to charged capacitors. We have included some protection circuitry, but large capacitors can pack enough of a punch to render that protection moot! The same advice applies to our ESR Test Tweezers.

Like Bob Parker's classic and now 20-year-old ESR Meter Mk2 design (https://zilogbob.com/esr_meter/esrmeter.htm), the ESR Test Tweezers are also well suited to measuring low resistances, such as current shunts. So they are sure to come in handy for other sorts of troubleshooting.

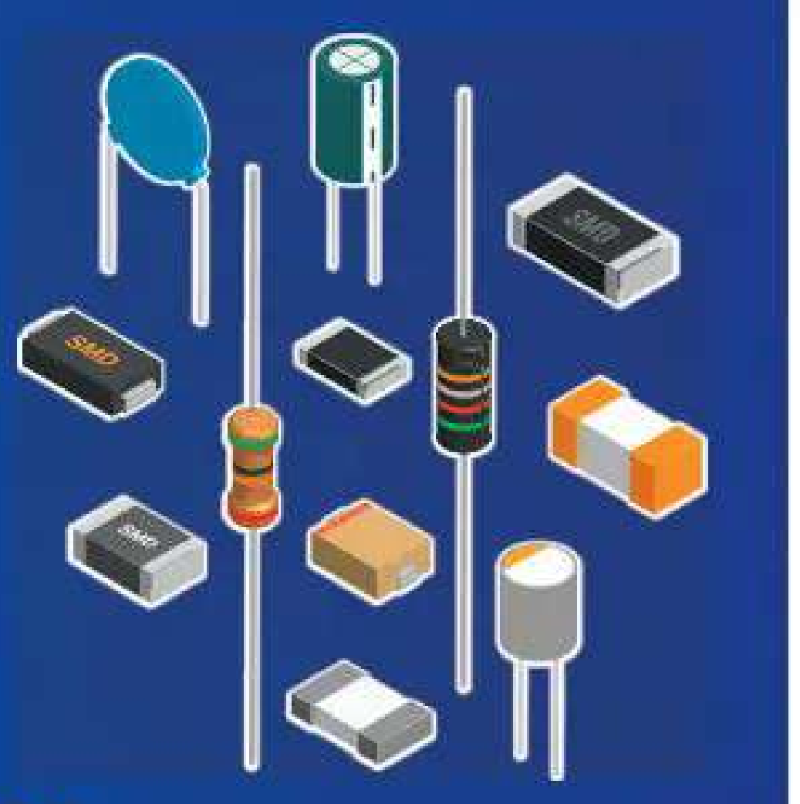
The Arduino-based LC and ESR Meter from August 2024 uses the same 'front-end' design as the ESR Meter Mk2 to measure ESR, but piggy-backs



ESR Test Tweezers

Features & Specifications

- ☒ Measures ESR/resistance from 0.01Ω to 1kΩ
- ☒ Measures capacitance from 100nF to 50µF
- ☒ Can perform in-circuit testing as long as capacitors are discharged
- ☒ Compact Tweezers format makes probing parts easy
- ☒ Runs from a single 3V lithium coin cell
- ☒ Will operate down to a cell voltage of 2.4V
- ☒ Displays results on a clearly visible OLED screen
- ☒ Typical accuracy better than 10%
- ☒ Adjustable sleep timeout and brightness
- ☒ Display can be rotated to suit left- and right-handed use
- ☒ Simple calibration of most parameters
- ☒ The standby cell life is close to the cell shelf life



Silicon Chip ESR Test Tweezers Kit (SC6952, ~£30 + P&P)
This kit includes everything in the parts list except the coin cell & optional header CON1. The three resistors & one capacitor needed for calibration are included.

onto the Wide-range digital LC Meter from the June 2019 issue, using its processor to drive the measurement circuitry and display the results.

That was a popular project, but we reckoned we could simplify the all-important ESR sensing circuitry and fit it into a much more compact instrument that costs less to build.

Measuring ESR

Measuring ESR is not difficult in theory, although we must be able to separate the effects of the main capacitance and leakage resistance from the ESR (see Fig.1). As we noted, the ESR is often taken to include ESL at a specific frequency, so we don't need to concern ourselves with ESL too much.

The ESR Test Tweezers use the same philosophy as the other ESR Meters. Relatively low currents are briefly pulsed into the capacitor, and the voltage across the capacitor is measured. It is allowed to discharge between tests.

The brief pulses do not have time to significantly charge the capacitor (assuming it is above 1µF); the capacitance acts like a short-circuit in this testing, so it does not affect the reading.

Since the capacitor is practically always discharged, the leakage resistance has no effect; the capacitance effectively short-circuits it. The pulses can also be considered analogous to an AC signal, so the capacitor's impedance is low enough that the ESR dominates.

Knowing the ESR is not enough to tell whether a capacitor is faulty. It's a good idea to verify that its capacitance hasn't dropped, and this Meter can do that, too, up to about 50µF. Beyond that, most DMMs will have a capacitance measurement mode that works up to a few thousand microfarads.

Any decent capacitor will specify its expected ESR value (or equivalent) in the data sheet, and you can compare that value to the Meter's reading. However, when servicing equipment, the exact part number may not be known, so the earlier ESR Meters provided a table showing roughly acceptable values for a range of capacitors.

Table 1 shows these values. Some data sheets might specify a dissipation factor or loss angle instead of an ESR value; page 47 has information about what those parameters mean and how to convert them to an ESR value.

The ESR Test Tweezers are much smaller than the earlier devices, so we have not been able to include the table on the equipment, but you can download it, print it out and keep a copy handy.

Design compromises

This device is patterned on the very popular Advanced Test Tweezers from February & March 2024. They are a compact and elegant device with many useful functions. So we have kept the ESR Test Tweezers to much the same form factor, using differently-coloured PCBs to make the two tools easier to tell apart. We know that many readers will end up with both!

The Advanced Test Tweezers performed most of the tests in software running on a microcontroller, so they needed relatively few external components. For testing ESR, we need more complicated circuitry, so we have had to use more components. They are the same M2012 (0805 imperial) SMD parts that measure 2.0 × 1.2mm along with a few other parts in small packages. Apart from there being more components, construction should not be any harder than for the Advanced SMD Tweezers.

The ESR Test Tweezers use simplified circuitry compared to the earlier

Table 1: typical ESR readings for good capacitors

	10V	16V	25V	35V	63V	160V	250V
1µF			5	4	6	10	20
2.2µF			2.5	3	4	9	14
4.7µF			6	3	2	6	5
10µF		1.6	1.5	1.7	2	3	6
22µF	3	0.8	2	1	0.8	1.6	3
47µF	1	2	1	1	0.6	1	2
100µF	0.6	0.9	0.5	0.5	0.3	0.5	1
220µF	0.3	0.4	0.4	0.2	0.15	0.25	0.5
470µF	0.15	0.2	0.25	0.1	0.1	0.2	0.3
1000µF	0.1	0.1	0.1	0.04	0.04	0.15	
4700µF	0.06	0.05	0.05	0.05	0.05		
10mF	0.04	0.03	0.03	0.03			

If your capacitor's data sheet does not mention a typical or maximum ESR value, this table can be used as a guide. If your data sheet mentions a dissipation factor or loss angle, refer to our panel on page 47. This table can be downloaded from siliconchip.com.au/Shop/11/238

ESR Meter designs. That's partly to help us fit the parts on the board but also because we were able to reduce the parts count without compromising performance, saving on parts cost and assembly time.

For example, the older designs feature a pulse injector with 11 parts and a pulse amplifier made from 17 parts. The corresponding sections of our circuit have only five and nine parts, respectively (50% less overall!). We are not using a voltage regulator either; instead, our software compensates for any variations in the supply voltage from the cell.

The earlier designs used a comparator (built into the processor) alongside a voltage ramp to measure the pulse amplitude, requiring eight more parts. Our circuit uses the 12-bit ADC (analog-to-digital converter) peripheral built into the microcontroller and no external parts.

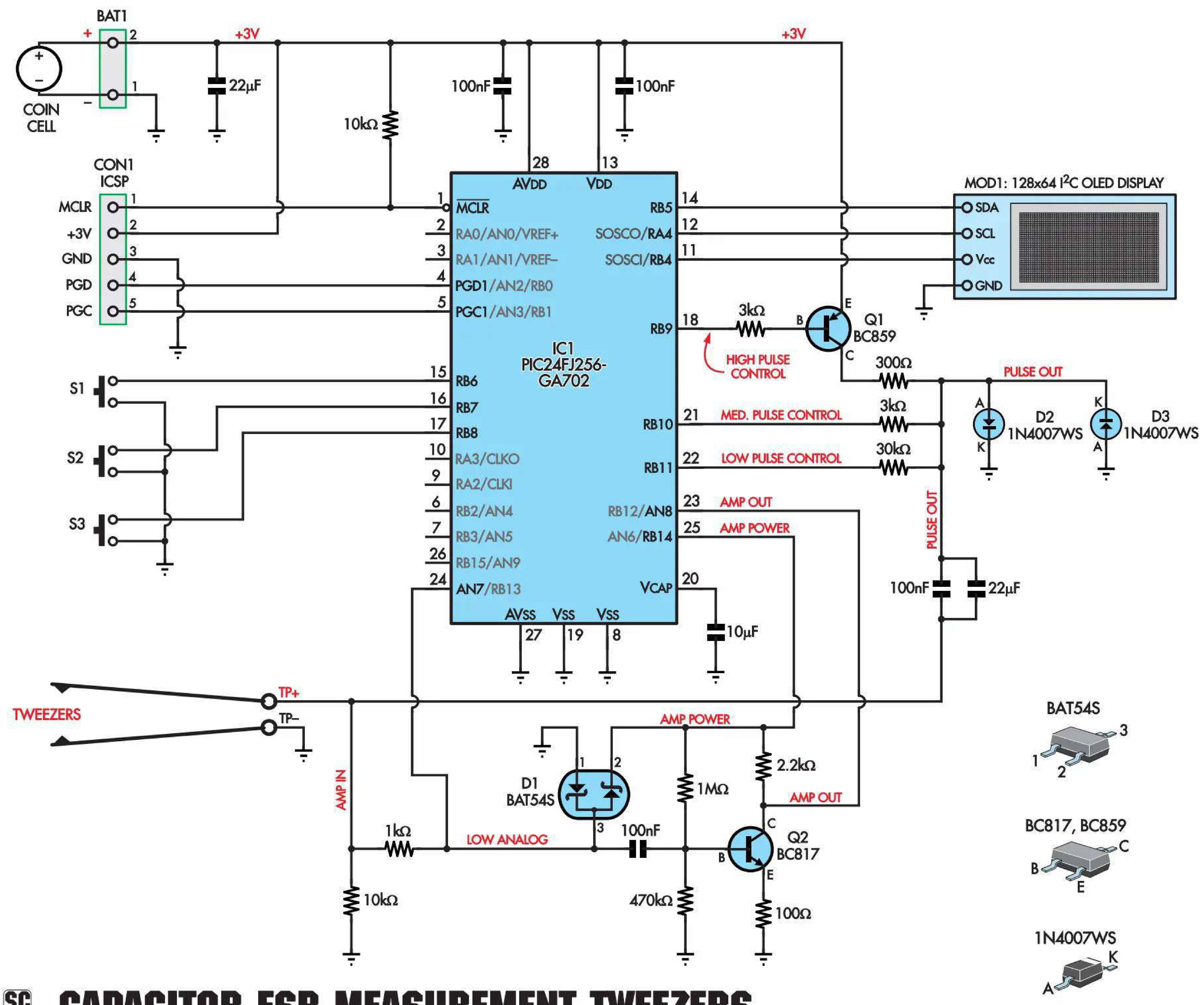
Instead of a multiplexed LED display driven by a shift register IC, requiring several more parts, we are using the same graphical OLED display module as in the Advanced Test Tweezers (although it's white this time rather than blue/cyan). It sits over the main PCB, occupying only the size of a four-pin header on the main PCB.

The earlier ESR Meters could apply test pulses up to 50mA. Given that the ESR Test Tweezers are designed to run from a coin cell, we aimed to use lower amplitude pulses to avoid excessive drain from the cell.

Despite all this, the ESR Test Tweezers can measure fairly accurately down to 10mΩ (just like our previous ESR meters) and will draw less than 1μA of current when in low-power mode; that's low enough that the standby life of the cell will be close to its shelf life.

We tested our prototype using our Coin Cell Emulator that was described in the November 2024 issue. It reported a current of 0.0μA while the ESR Test Tweezers were sleeping, less than the 100nA minimum that the Coin Cell Emulator can display.

The typical operating current is around 3.5mA with no components connected to the test leads, rising


BAT1
COIN CELL
+3V
22μF
10kΩ
100nF
100nF
CON1 ICSP
MCLR
+3V
GND
PGD
PGC
28 AVDD
13 VDD
IC1 PIC24FJ256-GA702
MCLR
RA0/AN0/VREF+
RA1/AN1/VREF–
PGD1/AN2/RB0
PGC1/AN3/RB1
RB6
RB7
RB8
RA3/CLKO
RA2/CLKI
RB2/AN4
RB3/AN5
RB15/AN9
AN7/RB13
RB5
SOSCO/RA4
SOSCI/RB4
RB9
RB10
RB11
RB12/AN8
AN6/RB14
VCAP
AVSS
VSS
VSS
10μF
MOD1: 128x64 I2C OLED DISPLAY
SDA
SCL
Vcc
GND
3kΩ
Q1 BC859
HIGH PULSE CONTROL
300Ω
PULSE OUT
3kΩ
MED. PULSE CONTROL
30kΩ
LOW PULSE CONTROL
D2 1N4007WS
D3 1N4007WS
AMP OUT
AMP POWER
PULSE OUT
100nF
22μF
S1
S2
S3
TWEEZERS
TP+
TP–
AMP IN
1kΩ
10kΩ
LOW ANALOG
D1 BAT54S
100nF
1MΩ
470kΩ
2.2kΩ
AMP OUT
Q2 BC817
100Ω
BAT54S
BC817, BC859
1N4007WS

SC ©2024 CAPACITOR ESR MEASUREMENT TWEEZERS

Fig.2: the ESR Test Tweezers use a 16-bit, 28-pin PIC24 microcontroller to drive the measurement circuitry and a small OLED display. Different test currents are applied to the DUT via the 300Ω, 3kΩ and 30kΩ resistors, while Q2 amplifies the voltage across it for the micro to sense using its internal ADC. The diodes protect the micro in case the probed capacitor has some charge left.

to 5mA when a component is being tested or settings are being modified. About half of that current is due to the OLED screen, which is set to near its lowest brightness setting by default.

The current draw increases if you need to operate the OLED at a higher brightness, but we found that was not necessary for indoor use.

Circuit details

Fig.2 shows the full circuit diagram of the ESR Test Tweezers. Many components are common to the Advanced Test Tweezers: IC1, MOD1 and CON1 are much the same, with IC1 being the PIC24FJ256GA702 16-bit microcontroller.

IC1 is powered by coin cell BAT1. The two 100nF capacitors bypass its two positive supply pins, while the 10μF capacitor provides bypassing for a 1.8V regulator internal to IC1. Practically nothing else is connected directly to the cell, meaning that IC1 has total control over what can draw current from it.

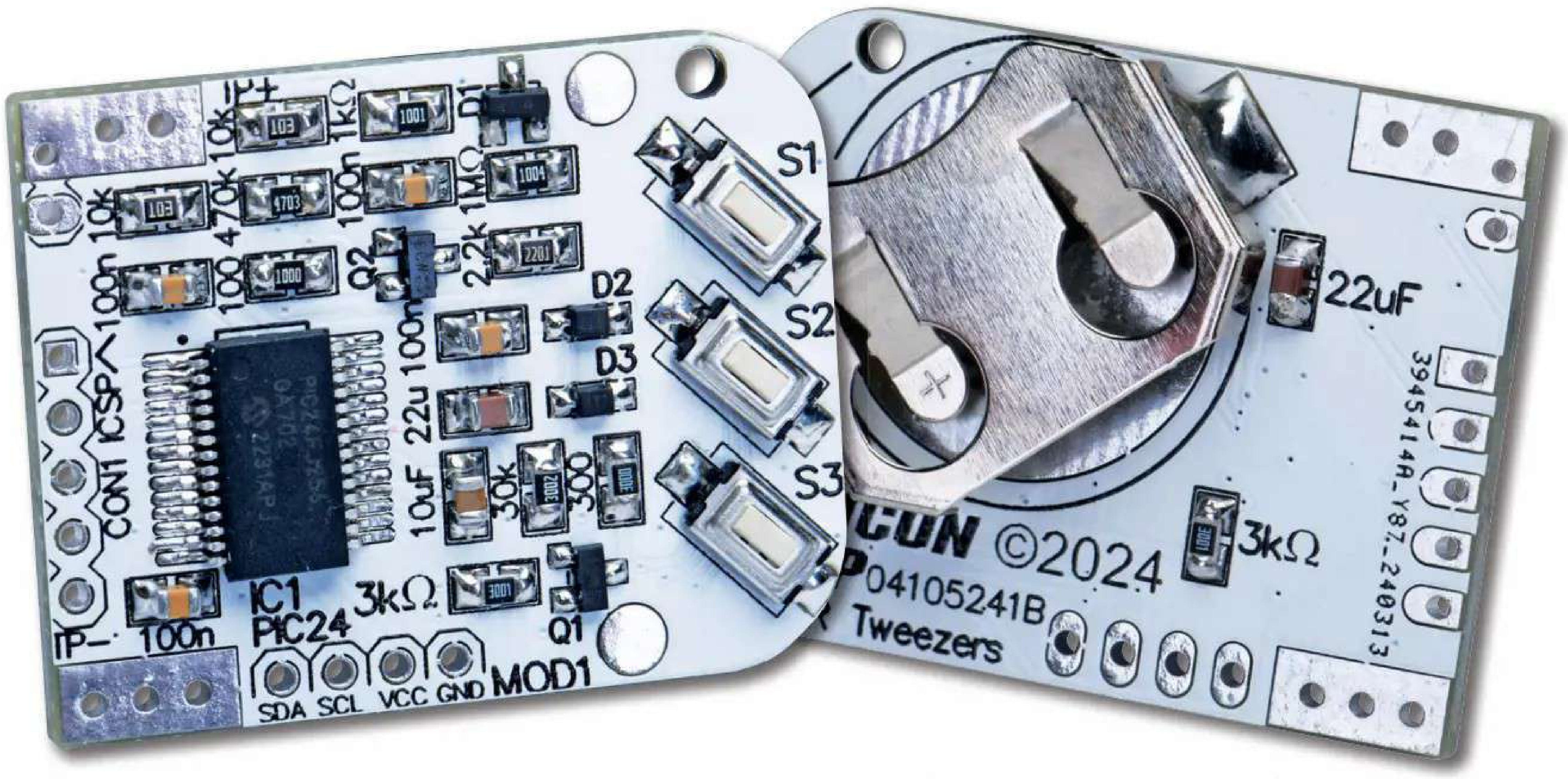
The 22μF capacitor provides a reserve of power to assist the coin cell in delivering the test pulse current. This is about the highest value of capacitor commonly available in the M2012 size we are using for this project; it is sufficient for our needs.

The highest pulse current is 10mA, applied for no more than 50μs. With a 22μF capacitor, the nominally 3V rail dips by about 0.02V, rather than the 0.2V expected without the capacitor. This also means that the coin cell is subjected to a lower average load; it does not see the heavy peaks that would otherwise shorten its useful life considerably.

CON1 is the ICSP (in-circuit serial programming) header and the 10kΩ resistor on IC1's pin 1 sets the micro to run normally unless a programmer is connected. We mainly included CON1 to simplify software development; you shouldn't need it in regular operation, although it may be useful if we ever release a firmware update.

MOD1 is an I²C OLED module powered at its V_{CC} pin by one of IC1's I/O (input/output) pins. Pulling that pin low shuts off the display module completely. The other two connected I/O pins provide the I²C serial control interface.

Tactile pushbuttons S1-S3 connect to three more I/O pins. Each is furnished with an internal pullup current from IC1, so their state can be easily detected without external parts. Debouncing is done by the software.



The ESR Test Tweezers PCB (shown enlarged) looks similar to the Advanced Test Tweezers, but it has different capabilities. We used white PCBs to set them apart and will provide white arm PCBs to match.

The parts below MOD1 form the pulse injection circuitry. The 300Ω, 3kΩ and 30kΩ resistors allow nominal currents of 10mA, 1mA and 100μA to be generated from a 3V supply rail.

IC1's I/O pins can source 1mA with only a small (less than 0.1V) voltage drop. At 10mA, the drop would be around 0.6V, so the 300Ω resistor is provided with PNP transistor Q1 for switching; the second 3kΩ resistor provides the base current when Q1 is driven.

The 22μF and 100nF capacitors in parallel are present to limit the amount of charge that can be injected if a large, charged capacitor is connected to the TP+ and TP− terminals. They act together as a low impedance when the pulses are applied. Silicon diodes D2 and D3 clamp any voltage from the capacitor being tested that exceeds their forward thresholds.

The presence of D2 and D3 also means that the maximum pulse that can be applied is less than 1V. So even if you test a capacitor in reverse, the voltage should be low enough to avoid damaging it.

IC1's pins 21 and pin 22 are normally kept low, and pin 18 is kept high, turning Q1 off. The PULSE OUT line sits at 0V and the 22μF and 100nF capacitors are discharged via the 10kΩ resistor at bottom left. Any connected device is also discharged.

Just before a pulse is applied, pins 21 and 22 are put in a high-impedance state by the processor. The appropriate pin is driven high (or low in the case of pin 18) to start the pulse. A measurement is then taken, and the pins revert to their idle state, ready for the next measurement.

Sense amplifier

The DUT (device under test), usually a capacitor or low-value resistor, connects between the TP+ and TP− pins. The test current applied to the PULSE OUT line induces a voltage at TP+ relative to circuit ground. The circuitry below IC1 amplifies the resulting voltage. When IC1's pin 25 is low, this circuitry is powered off via the AMP POWER line, but it is brought high during testing.

The 1MΩ/470kΩ divider ensures that Q2 is biased on slightly, as long as the supply is above about 2V. The 100nF capacitor at Q2's base will have the bias voltage across it. Before a pulse is applied, the voltages at LOW ANALOG (pin 24, AN7) and AMP OUT (pin 23, AN8) can be sampled by IC1's ADC to record a baseline voltage.

The LOW ANALOG line will be close to 0V, and the AMP OUT pin will be close to the voltage provided by the AMP POWER line, which will be reduced slightly due to Q2 being biased on slightly.

When a pulse is applied, the voltage rises at the TP+ pin, and the voltage at Q2's base rises by a similar but slightly smaller amount. The reduction is due to the signal being attenuated by the surrounding components, such as the 10kΩ resistor and 1MΩ/470kΩ divider.

Q2 behaves as an emitter follower, so its emitter will rise by much the same voltage, and the current through the 100Ω resistor will be proportional to the emitter voltage.

Since the collector current will match the emitter current (give or take the much smaller base current), the current through the 2.2kΩ resistor will be the same as that through the 100Ω resistor, meaning that the voltage across the 2.2kΩ resistor is 22 times that across the 100Ω resistor.

The microcontroller then takes another sample to compare with the baseline values. In practice, the change at the AMP OUT pin is 10-15 times the change at the LOW ANALOG line. Of course, the AMP OUT line will fall during a pulse, while the LOW ANALOG line will rise, but it is simple enough to take the difference either way.

The 1kΩ resistor and dual diode D1 provide another level of protection against external voltage sources (such as charged capacitors).

While it appears that we effectively have six ranges to read (two analog inputs multiplied by three current sources), they overlap. We use four ranges: the 100µA source sensed at the LOW ANALOG input and all three test currents sensed at the AMP OUT input.

Note that neither the LOW ANALOG or AMP OUT signals can swing rail-to-rail. Diode D1 clamps the LOW ANALOG level between AMP POWER and ground. Due to the 100Ω resistor, the AMP OUT signal cannot reach 0V, even if Q2 is saturated.

Several calibration factors are programmed into the ESR Test Tweezers, including the levels at which the LOW ANALOG and AMP OUT signals are valid.

Firmware

The firmware driving the ESR Test Tweezers has much in common with the Advanced Test Tweezers since they use the same microcontroller. However, the ESR Test Tweezers do not have as many features.

We have implemented three measurement modes, labelled ESR, RES and CAP. The ESR mode provides a function similar to our previous ESR meters.

The main ESR testing mode uses the 100µA source and the LOW ANALOG input to detect if a component is present across TP+ and TP−. If so, it runs pulses from each of the 100µA, 1mA and 10mA sources, taking measurements using the AMP OUT signal from the pulse amplifier.

Parts List – ESR Test Tweezers

1 double-sided main PCB coded 04105241, white solder mask, 36 × 28mm
2 double-sided arm PCBs coded 04106212, white solder mask, 100 × 8mm
1 double-sided back panel PCB coded 04105242, white solder mask, 36 × 28mm
1 0.96in 128×64 I^{2}C OLED module, white (MOD1)
1 surface-mounting 32mm coin cell holder (BAT1)
3 SMD two-pin tactile switches (S1-S3)
1 3-pin gold-plated header, 2.54mm pitch (for tips and mounting MOD1)
1 4-pin header, 2.54mm pitch (to mount MOD1; usually comes with MOD1)
1 5-way header, 2.54mm pitch (CON1; **optional**, for ICSP)
1 M2 × 6mm Nylon panhead machine screw
2 M2 Nylon hex nuts
1 CR2032 or CR2025 lithium coin cell
1 small piece (eg, 2 × 2cm) of double-sided foam-core tape
2 100mm lengths of 10mm diameter clear heatshrink tubing

Semiconductors
1 PIC24FJ256GA702-I/SS microcontroller programmed with 0410524A.HEX, SSOP-28 (IC1)
1 BC859 PNP transistor, SOT-23 (Q1; marking 4C)
1 BC817 NPN transistor, SOT-23 (Q2; marking 6C)
1 BAT54S dual schottky diode, SOT-23 (D1; marking KL4)
2 1N4007WS silicon diodes, SOD-323 (D2, D3)

Capacitors *(all SMD M2012/0805 size 6.3V+, X5R or X7R)*
2 22µF
1 10µF
4 100nF 50V X7R
extra 10µF (could be any type) for capacitance calibration

Resistors *(all SMD M2012/0805 size, $\frac{1}{8}$W, 1% – codes in brackets)*
1 1MΩ (105 or 1004)
1 470kΩ (474 or 4703)
1 30kΩ (303 or 3002)
2 10kΩ (103 or 1002)
2 3kΩ (302 or 3001)
1 2.2kΩ (222 or 2201)
1 1kΩ (102 or 1001)
1 300Ω (301 or 300R)
1 100Ω (101 or 100R)
extra 10Ω, 100Ω and 1kΩ resistors for calibration

If the 10mA pulse gives a valid AMP OUT reading, an ESR value is calculated using this data and a calibration factor. The 1mA pulse is checked next; if this is not valid, the ESR reading is taken from the 100µA pulse.

You can tell which range has been used from the number of decimal places displayed. The 10mA pulse gives a result to two decimal places (0.01Ω), while the 1mA pulse gives a result to the nearest tenth of an ohm and so on.

The RES mode (for resistance) is intended to measure the values of resistors, and it does so using only the 100µA source. That makes it a bit easier on the cell since there are no high-current pulses. The resolution of the RES mode is only around 10Ω; we expect it to be useful if you have many parts to sort through.

The CAP mode gives a reading for both capacitance and ESR for the device under test. It also uses the 100µA source but applies it for long enough to charge up the capacitor, although this is somewhat limited by the 22µF capacitance in series with the DUT.

It takes readings at 40µs, 400µs and 4ms from the start of the pulse. Our prototype gave us fairly accurate readings up to 50µF, so we've specified that as the maximum. The display will show dashes if the measured capacitance is higher than 50µF.

The lower limit of 100nF is due to the resolution being about 10nF; the readings will tend to be inaccurate below 100nF. Since we have collected much the same data as the RES mode, an ESR reading is given too, with the same limitations as that mode.

The firmware is also responsible for monitoring button presses and putting the processor to sleep when the device is not being used. There is a SETTINGS mode where preferences and calibration parameters can be changed, including the option to save the calibration and settings to flash memory.

We'll delve into the calibration, setup & operation of the ESR Test Tweezers once construction is complete.

Construction

The SSOP-package microcontroller and M2012 parts mean assembly is not overly difficult, but it best suits constructors with some experience working with SMDs. If you have built the Advanced Test Tweezers, you should have little trouble with the ESR Test Tweezers.

You will need a fine-tipped soldering iron, solder, flux paste and solder-wicking braid. You should also have a magnifier, SMD tweezers and a means of holding the PCB in place, such as Blu-Tack. Good lighting is highly recommended, along with fume extraction (or work outdoors or near a large open window).

Start by placing a little flux paste on the PCB pads for IC1 and rest it in place, checking that the pin 1 dot is in the correct position. Looking at the PCB with CON1 at the bottom, the text on the chip should be right-way-up. Check your build against the Fig.3 overlay diagram and accompanying photos.

Note that our photos show CON1 fitted (which isn't necessary unless you need to program the chip onboard). We also fitted a socket for MOD1 so we could remove the OLED if necessary; you can hard solder it using a standard pin header.

Tack solder a couple of IC1's leads and check that the other pins on both sides are correctly aligned. Adjust it if needed before carefully soldering the remaining pins. When finished, clean away any flux residue (eg, using alcohol) and closely inspect the soldering before proceeding, as it will be much easier to correct problems you find before more components are fitted.

If you have bridged any of the pins of the IC, add a dab of flux paste on top and then use solder-wicking braid to clear it. Verify that all pins have had solder flow onto both the pin and the pad; if it's just on the pin, it will not make a good connection to the PCB.

Fit the three SOT-23 devices next, being careful not to mix them up. Dual diode D1 is near the top of the PCB, with PNP transistor Q1 near the bottom. Q2, the NPN transistor, is near IC1. If you aren't sure which is which, they should have codes printed on the top. The parts list has likely codes (although they can vary by manufacturer).

In each case, apply a little flux paste to the pads, tack one lead, then check that the other two leads are within their pads before soldering them.

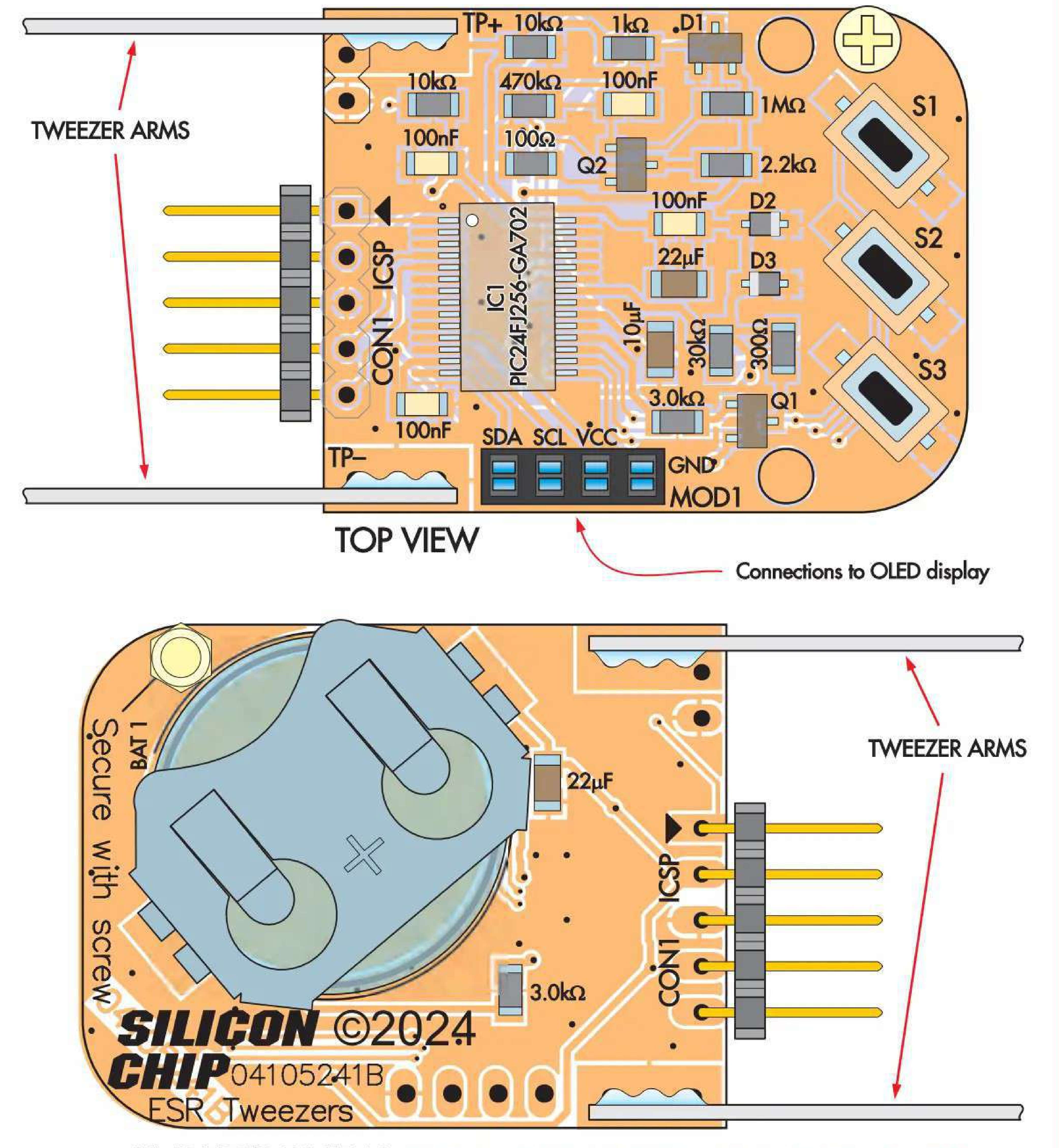
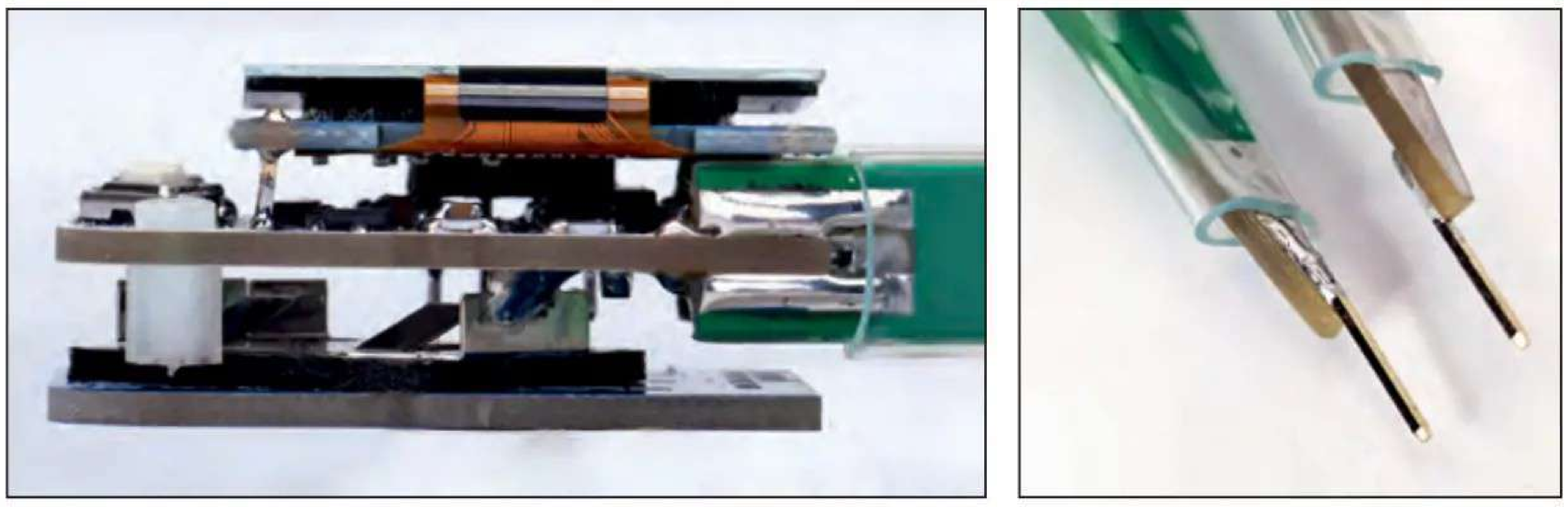


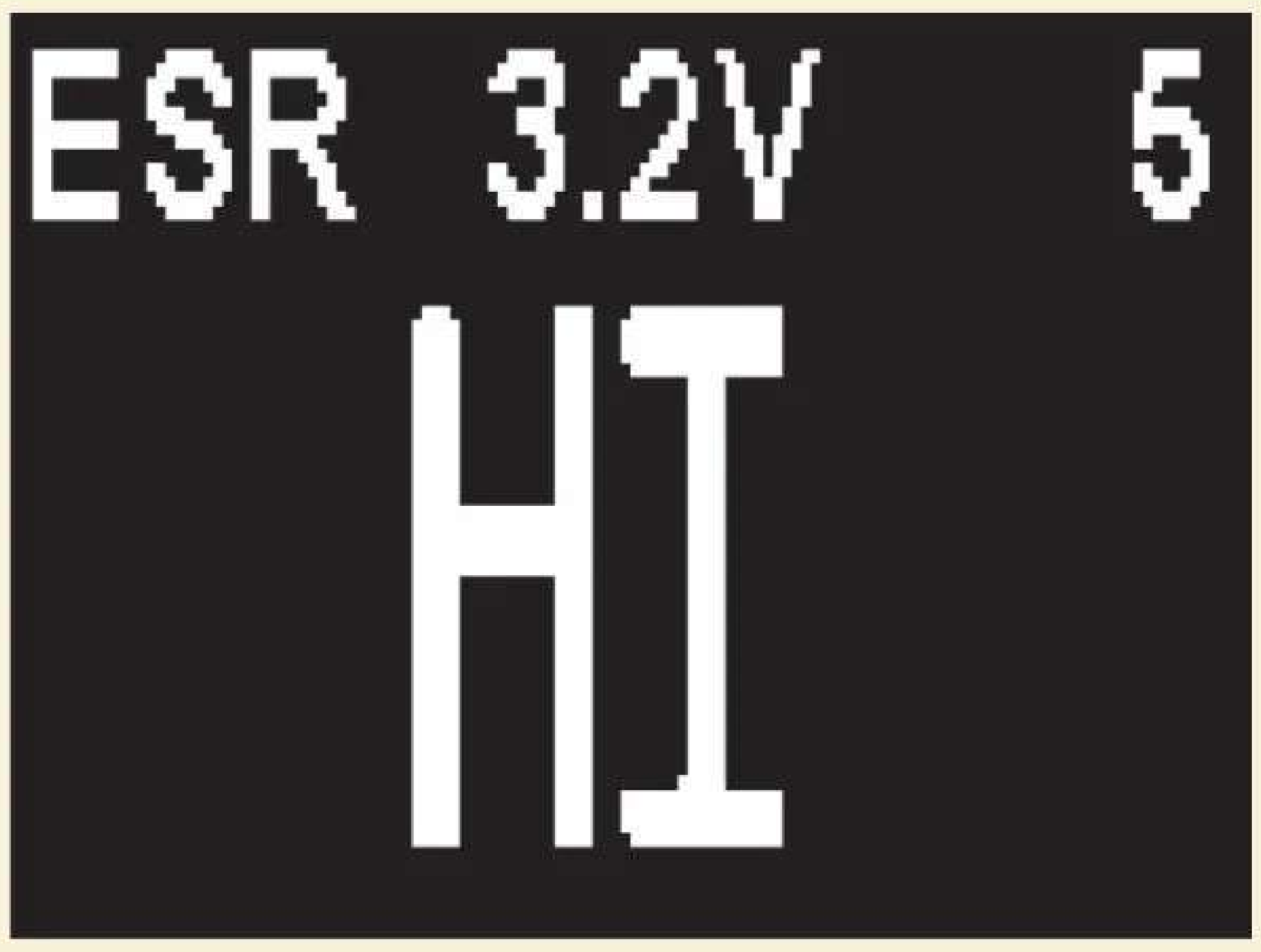
Fig.3: fit the components to both sides of the main PCB as shown here. Most of them are moderately easy to solder apart from IC1, which has closely spaced pins. Don't mix up the different SOT-23 devices and note that D2 and D3 are connected in opposite directions. You don't need to fit the headers for CON1 and MOD1; we did so to simplify the development process.



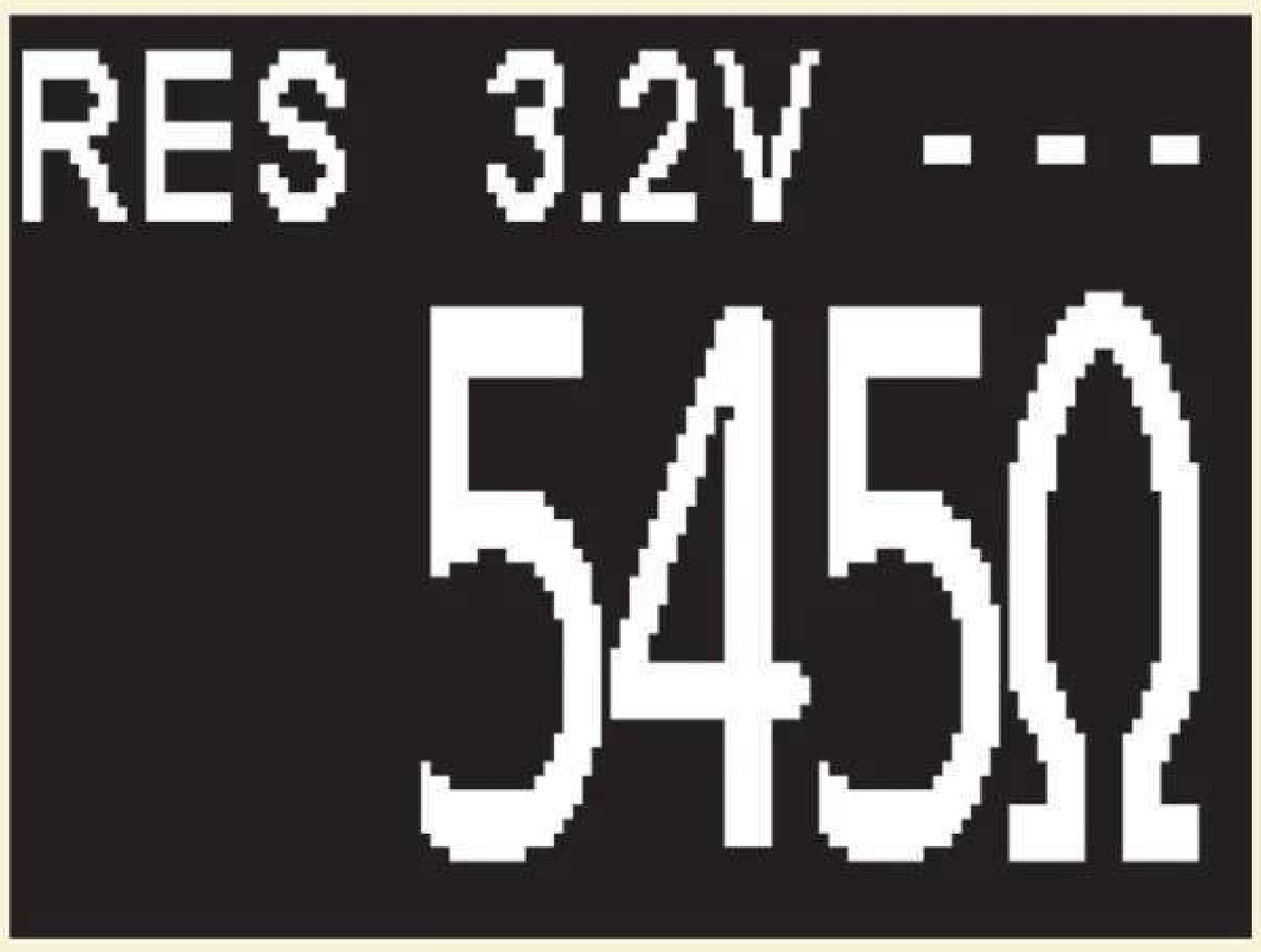
These photos show a number of the important construction details. The arms attach to the main PCB with chunky solder fillets and are protected by heatshrink tubing. The white screw and nuts prevent the coin cell from being easily removed. A header pin soldered between the main PCB and the OLED PCB helps to reinforce the OLED mounting. A solder fillet mechanically secures the tips to the arms. Ensure that the solder surrounds one end of the header pin and flows into the holes in the arm PCB.

The two single diodes, D2 and D3, face in opposite directions, so check that the PCB's cathode markings match the devices' cathode stripes.

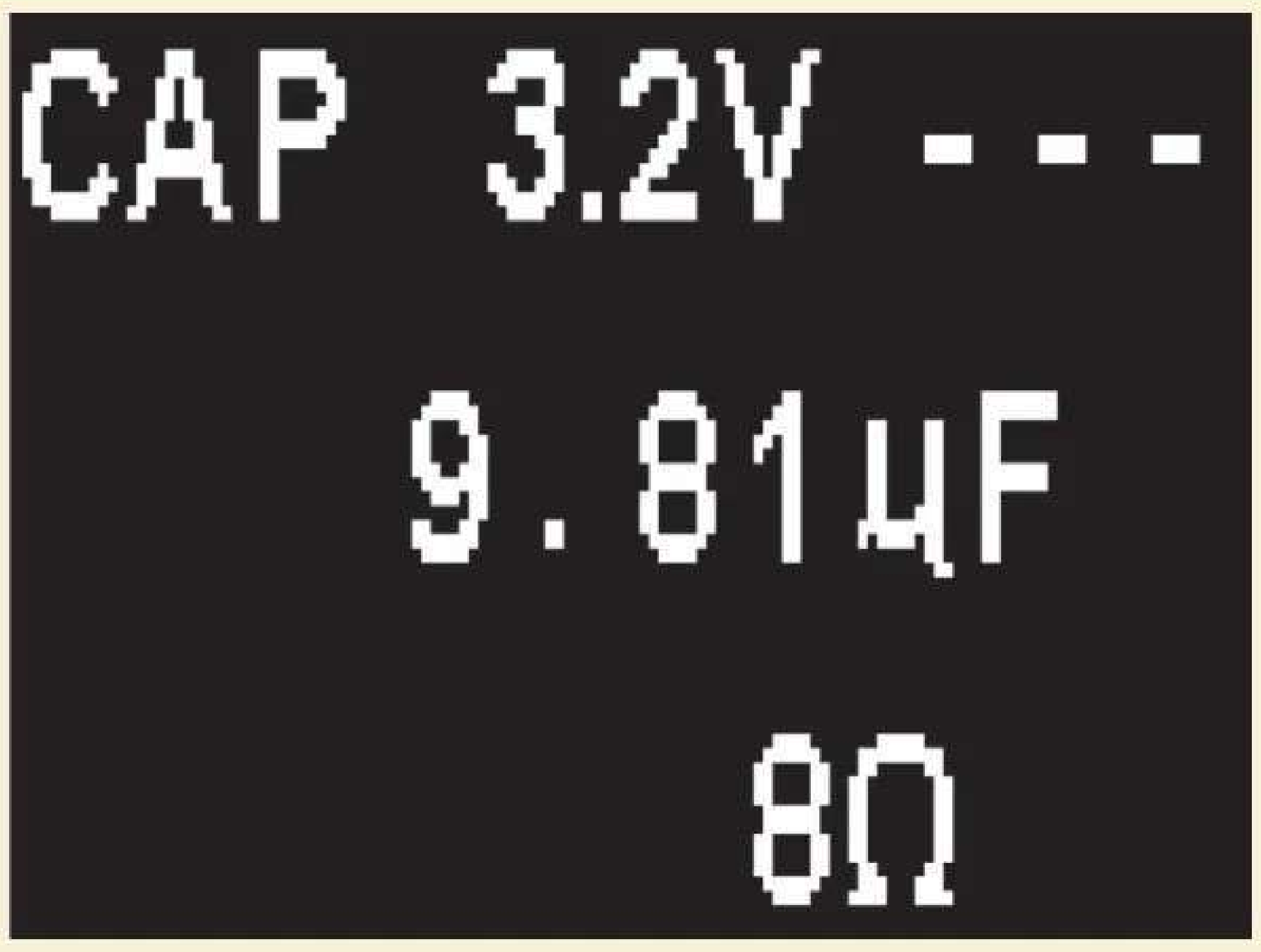
Fit the capacitors next, being careful not to mix them up, as they are not marked. There are four 100nF capacitors on the front of the PCB plus one 10µF capacitor. One of the 22µF capacitors is on the front, while the



Screen 1: the default display at power-on. Touching the tips together will show a low readings in ohms. The cell voltage is displayed next to a countdown timer; when the timer expires, the Tweezers enter a low-power sleep mode.



Screen 2: the second operating mode uses the low-current range to measure resistance without unnecessarily loading the cell. If S1 is pressed in any operating mode, the timer is paused and dashes are displayed, as seen here.



Screen 3: the third mode gives readings for capacitance (between 100nF and 50µF) and ESR using low-current pulses. A typical 10µF capacitor is connected here. Pressing S2 will resume the timer, as will changing modes with S3.

other mounts on the back of the PCB.

Now carefully work through the 11 resistors, matching the markings to the PCB silkscreen. The parts list shows the typical markings for the values we are using. Note that one of the 3kΩ parts is also on the back of the PCB.

Next, solder the cell holder to the back of the PCB. Make sure that the opening faces towards the screw hole; you can compare it to our photos.

Now thoroughly clean the flux residue off the PCB using a suitable solvent. Your flux might recommend one on its data sheet, but isopropyl alcohol is a good all-round alternative. Methylated spirits can be used, although it might leave residue. Allow the PCB to dry and inspect it again before proceeding.

Next, solder the three tactile switches, S1-S3. We do this now to avoid getting solvent in their mechanisms. They are fitted in much the same way as the other surface mounting parts but are a bit larger and easier to manage.

You can carefully clean up any flux residue from this step using a cotton tip or similar moistened by a small amount of solvent.

Programming the microcontroller

You won't need to perform this step if you have a pre-programmed microcontroller from the Silicon Chip Online Shop (including the one in our kit). If you have a blank micro, it's best to program it now before the arms and display are fitted, as they might get in the way.

You'll need a Snap, PICkit 3, PICkit 4 or PICkit 5 programmer to program the PIC24FJ256GA702 microcontroller. The Snap cannot provide power, so you can temporarily fit the coin cell while programming occurs.

We suggest using Microchip's free MPLAB X IPE for programming. It's available as part of the MPLAB X IDE download and can be installed on Windows, Mac and Linux computers. Choose the PIC24FJ256GA702 and open the 0410524A.HEX file in the IPE. Enable power from the programmer if you need it.

To avoid permanently soldering the header to the PCB, you can push the 5-way header into the socket on your programmer while holding the other ends of the pins in place through the pads of CON1. It's a bit of a juggle, but it will make the Tweezers easier to use later.

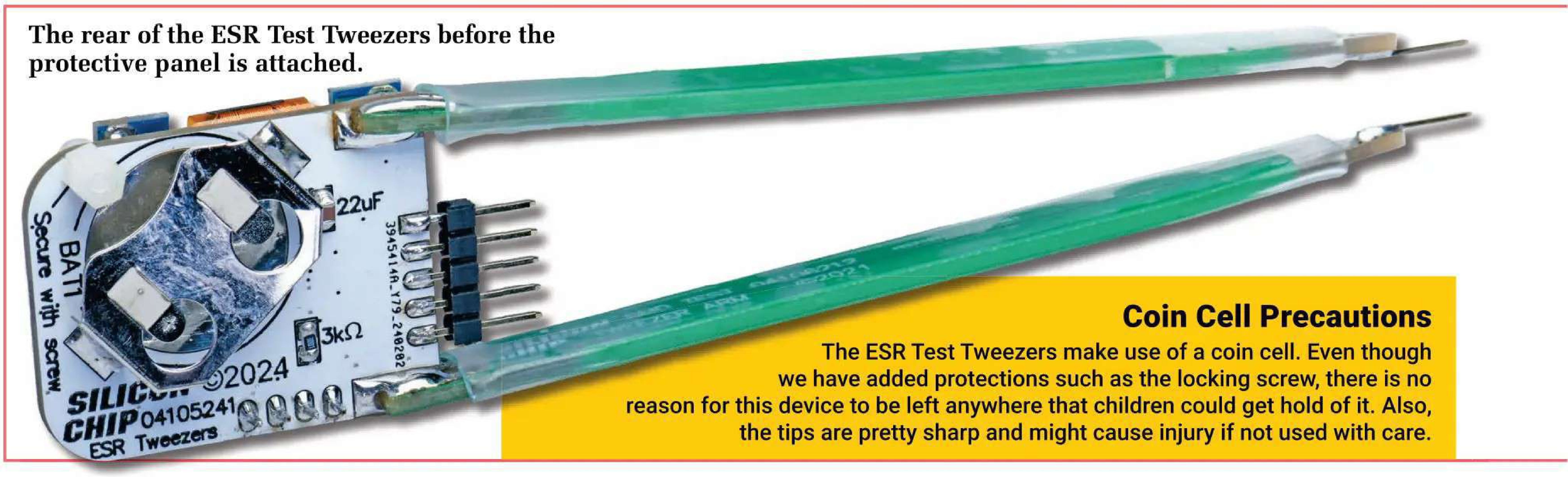
Click the button to program the chip and check that the IPE verifies the program correctly.

Fitting the arms

The arms are each formed from a long, thin PCB, with the tips using gold-plated header pins to offer a low-resistance contact surface that will not corrode. Tin each arm tip generously and remove the header pins from their shroud.

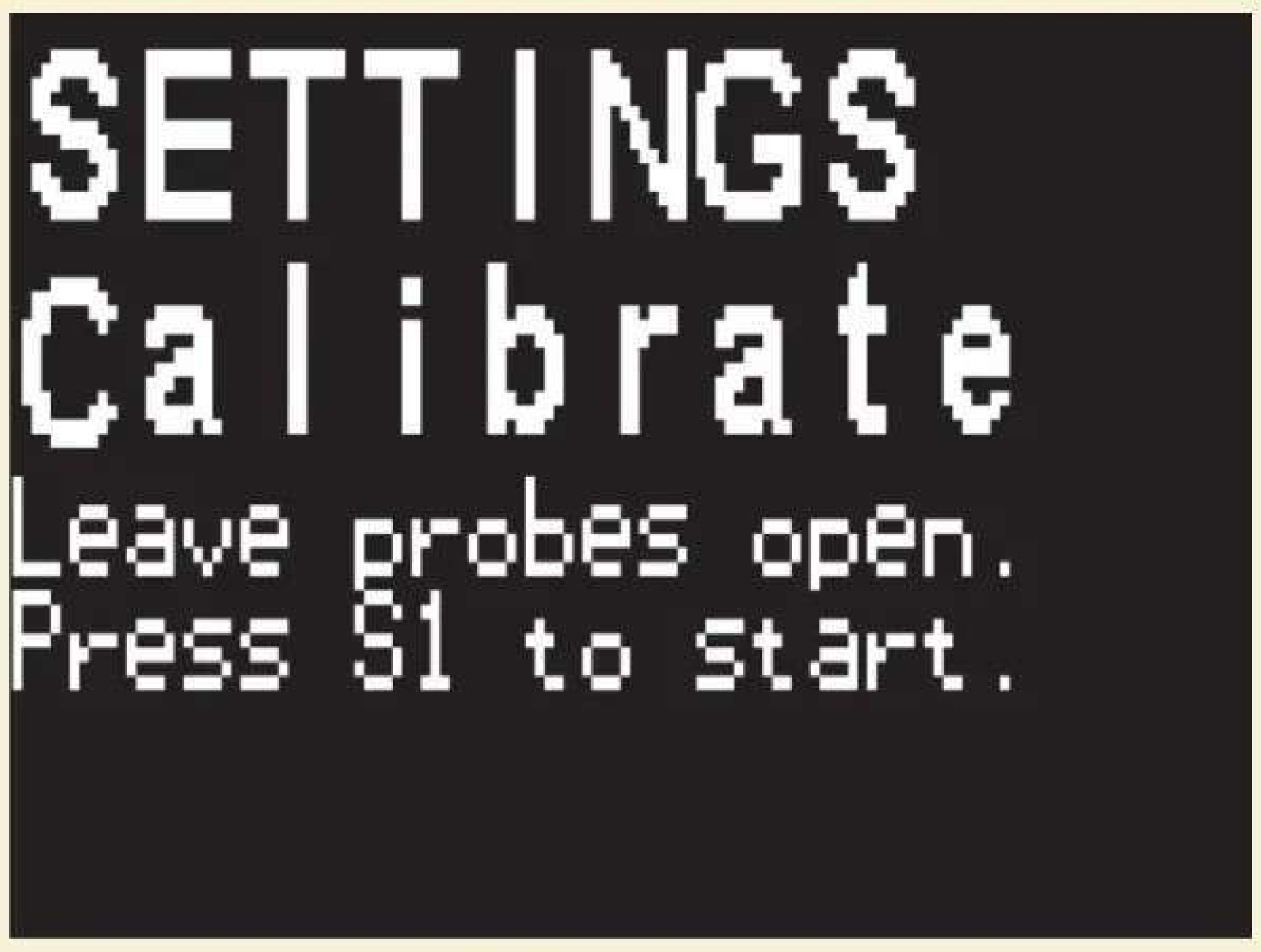
Using a pair of tweezers, solder a pin in position to the end of each arm, as shown in the photos. Try to line them up so they are centred. Note that the

The rear of the ESR Test Tweezers before the protective panel is attached.

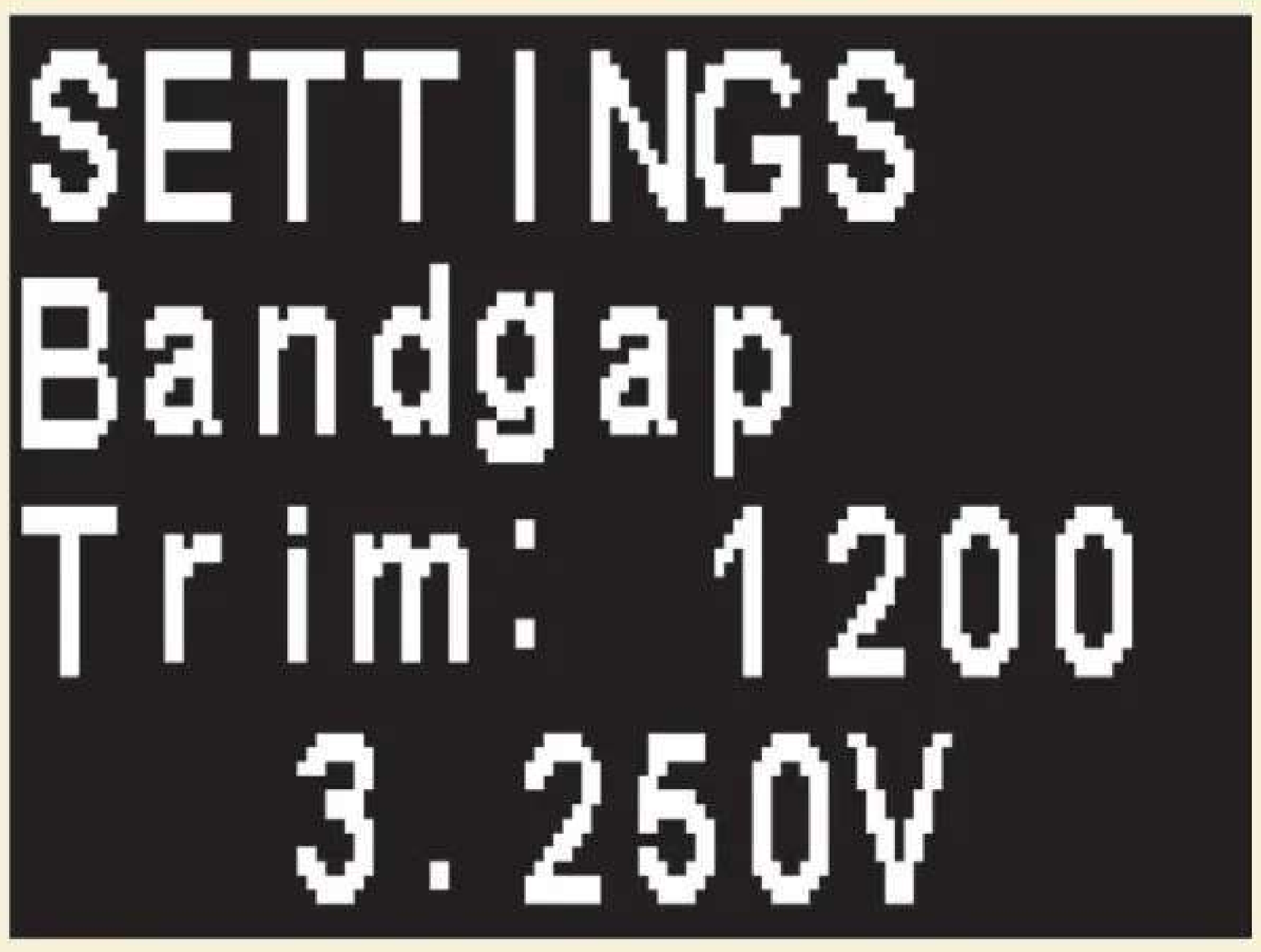


Coin Cell Precautions

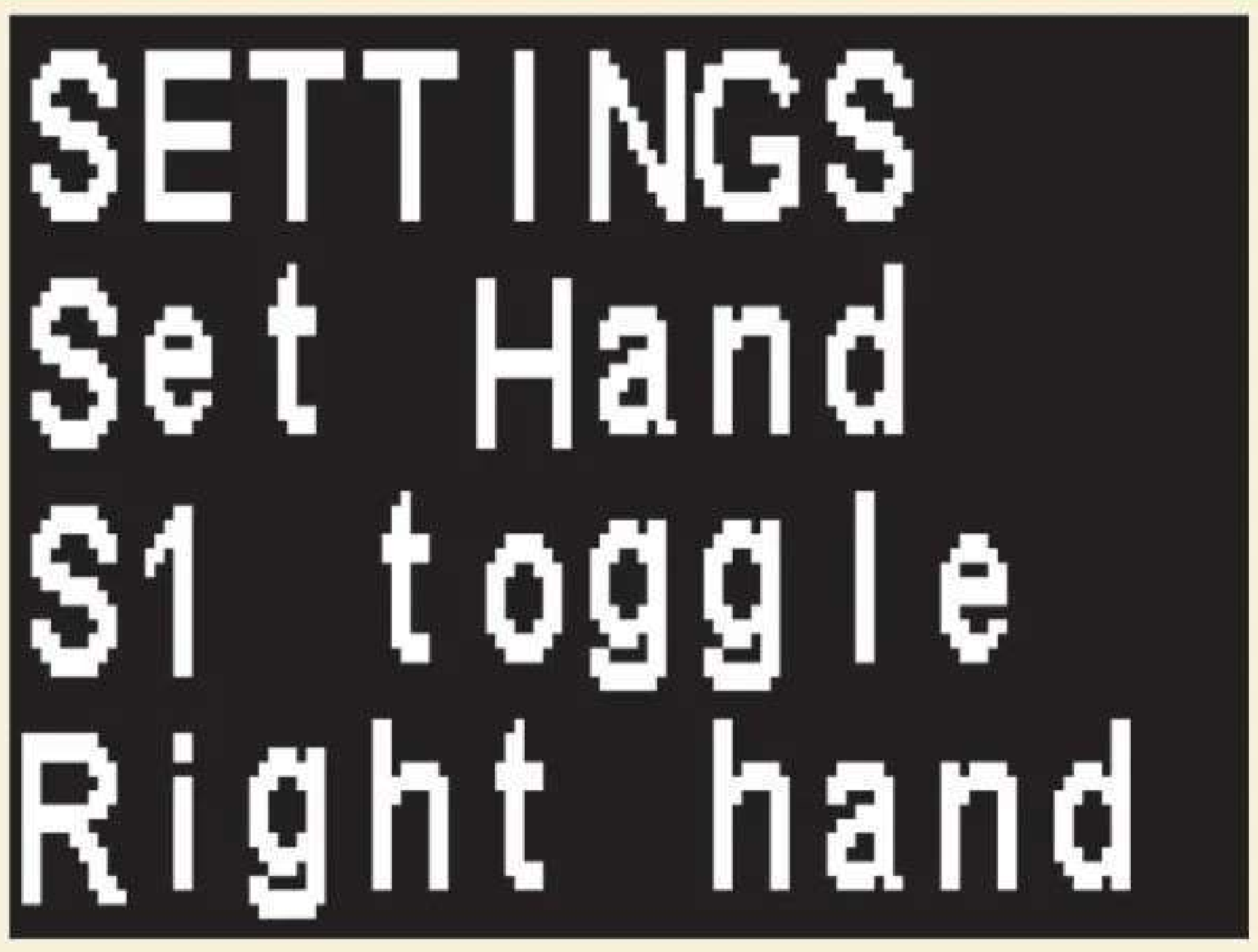
The ESR Test Tweezers make use of a coin cell. Even though we have added protections such as the locking screw, there is no reason for this device to be left anywhere that children could get hold of it. Also, the tips are pretty sharp and might cause injury if not used with care.


SETTINGS
Calibrate
Leave probes open.
Press S1 to start.

Screen 4: the Calibrate step takes readings with open and shorted tips and automatically sets the ADC saturation settings and probe (contact) resistance. Leave the tips open, press S1, then hold the tips together and press S1 again. Then release the tips. You can try again if you get an error.


SETTINGS
Bandgap
Trim: 1200
3.250V

Screen 5: the bandgap voltage is the nominally 1.2V reference used by IC1 for voltage measurements. At the bottom is the calculated supply (cell) voltage; use S1 & S2 to trim the bandgap until the displayed voltage matches the cell voltage, measured using a multimeter or similar.


SETTINGS
Set Hand
S1 toggle
Right hand

Screen 6: the display can be rotated by 180° to suit left- or right-handed use. Press S1 to toggle it and the display will rotate immediately to the new setting. Like all the other settings here, these new values are used immediately but are not automatically saved to non-volatile flash memory.

pins face inwards once the Tweezers are assembled.

The arm PCBs slot over the larger pads in the corners of the main PCB. We recommend not fully pushing the main PCB into the slot; leave some room. Take care that the arms do not contact any other pads on the PCB.

Fitting the arms is a bit like fitting the SMD components. Tack them roughly in place and check that they are aligned well, then add more solder to secure them firmly. Check the action and see that the tips meet correctly. Finally, add solid fillets of solder all-round to make them mechanically secure.

Slide the heatshrink tubing over the arms, leaving the tips clear, then shrink it in place. Doing this now avoids damage to the OLED screen from excessive heat.

We've taken some photos of the ends of the arms so you can see how the tips are attached and how the arms mount to the main PCB.

The OLED screen

The OLED is mounted next. You should be able to simply slot the four-way pin header into the pads of the MOD1 footprint on the PCB. We recommend temporarily placing a piece of card behind the OLED to prevent it from shorting the main PCB or arms. This will also help to add a small space between them. Tack one pin and check that the display is neat and square.

Solder the remaining pins and remove the piece of card. You can fit the battery at this stage and check that everything works. You should see something like Screen 1 when it is first powered on. The reading should show a low value (under 0.1Ω) when the tips are shorted together.

Remove the battery and solder a pin header or piece of solid wire to the top right corner of MOD1 and through to the main PCB underneath. This provides extra support for the OLED to prevent it from flexing and touching the main PCB.

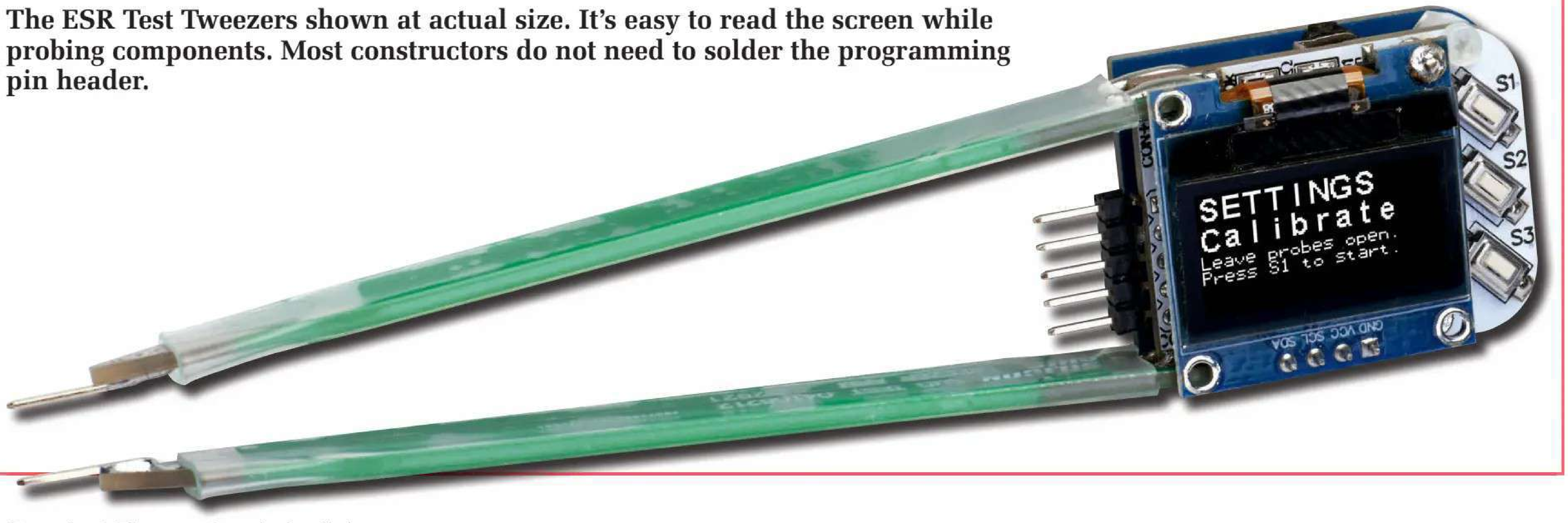
The back panel PCB can be soldered to the ground pins of MOD1 and CON1 or simply stuck to the back of the cell holder using double-sided tape. Ensure that the ESR TWEEZERS legend faces outwards (it's a dual-use panel; the other side has the legend for the Advanced Test Tweezers).

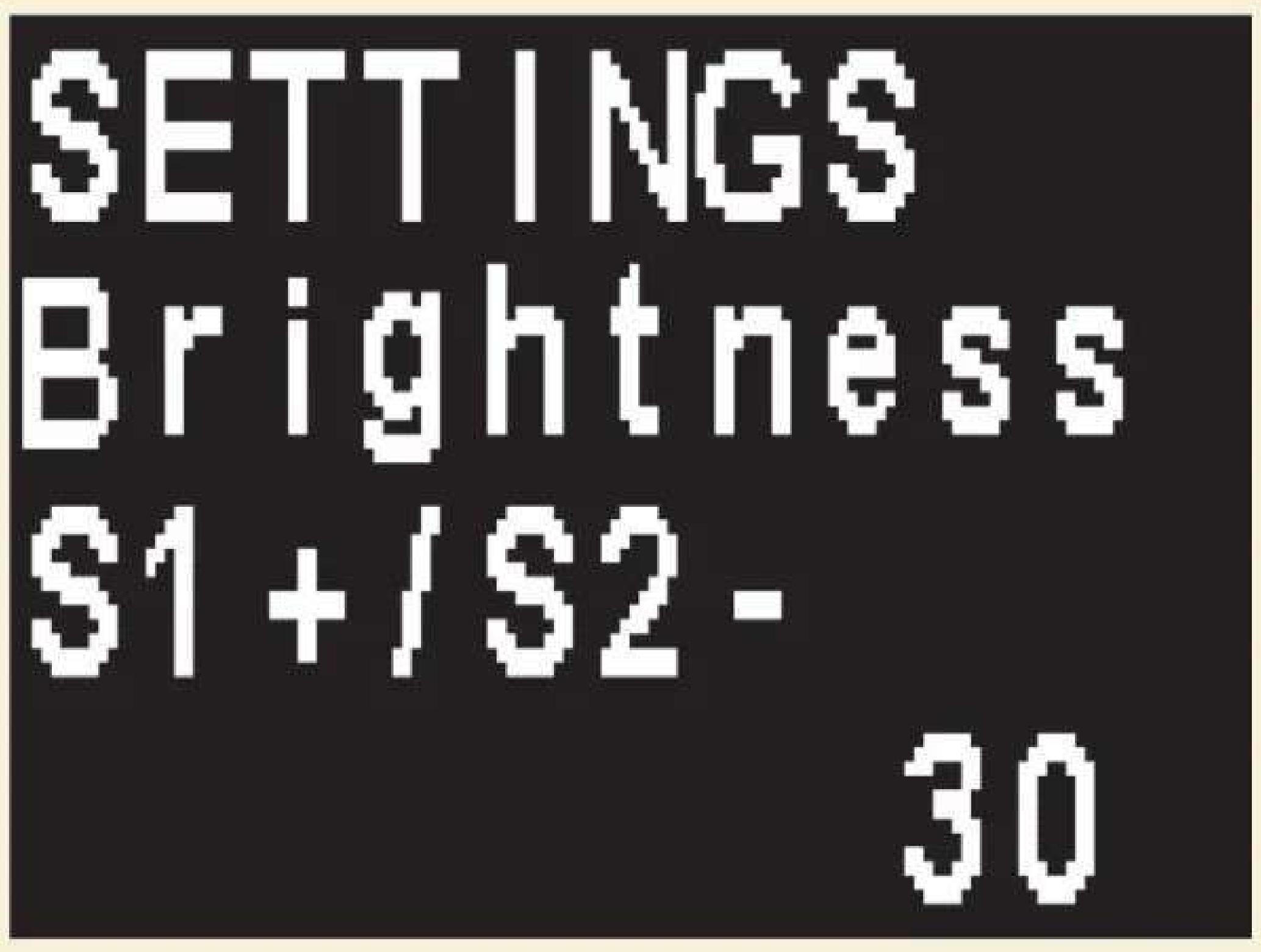
Finally, fit and secure the cell using the M2 Nylon screw and nuts. The nuts go on the same side as the cell, giving the depth needed to prevent the cell from being easily pulled out. The photos show how we have done that on our prototype. This is to prevent a child who might get hold of the Tweezers from removing the cell, which could be dangerous (it is hard to pull out regardless, but this is worthwhile extra security).

Calibration and operation

In regular operation, pushbutton S3 cycles between the modes, while S1 pauses the countdown timer. S2 (or any S3 mode change) will enable it again. The timer is shown at upper

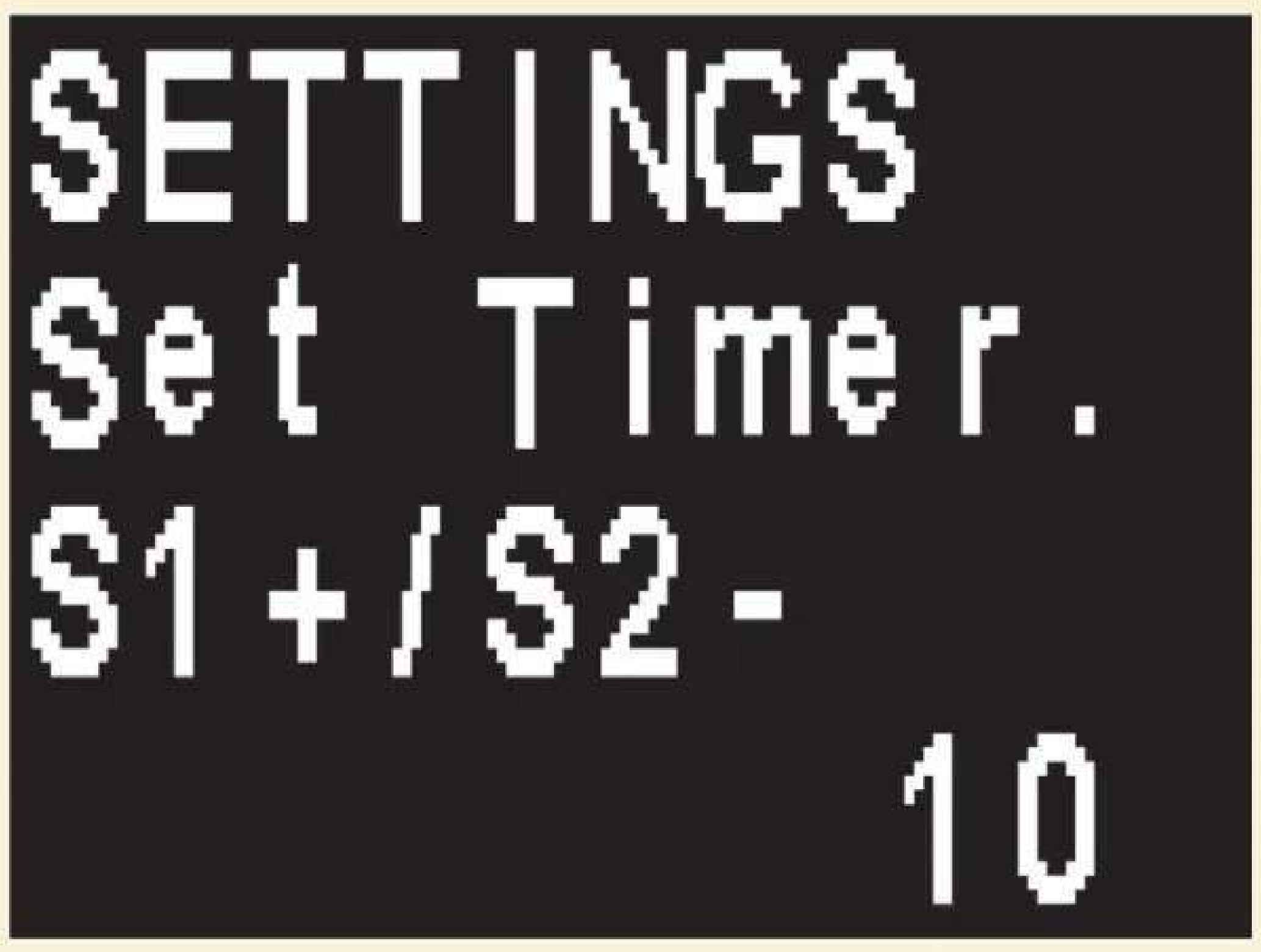
The ESR Test Tweezers shown at actual size. It's easy to read the screen while probing components. Most constructors do not need to solder the programming pin header.





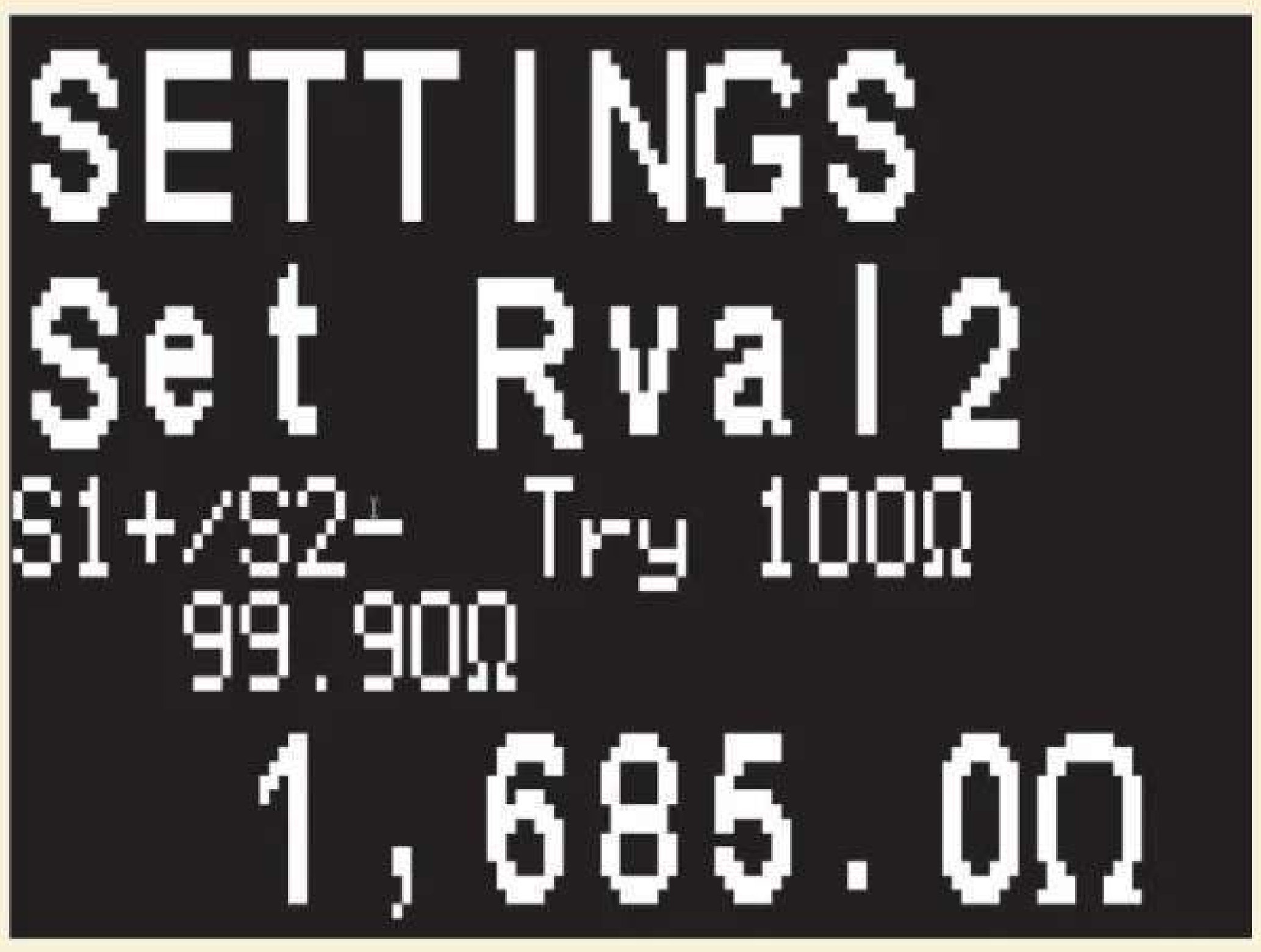
SETTINGS
Brightness
S1+/S2-
30

Screen 7: as with our other Tweezers, the OLED current draw is the single most significant drain on the cell. Setting the display brightness as low as possible (using S1 & S2) will prolong the cell life. The default level of 30 is the lowest usable setting; it can be changed in steps of five up to 255.



SETTINGS
Set Timer.
S1+/S2-
10

Screen 8: the timer is displayed in the ESR, RES and CAP modes. The Tweezers go into a low-power sleep when it counts down to zero. The time can be set in multiples of five seconds up to 995 seconds (about 16 minutes). Since the timer can be paused, you might not need to change this setting.



SETTINGS
Set Rval2
S1+/S2- Try 100Ω
99.90Ω
1,685.0Ω

Screen 9: four screens like this calibrate the current pulse values. Connect the recommended resistor or capacitor value (100Ω here) across the probes and trim the value until the smaller text (99.90Ω) is close to the actual value connected. The default values are based on our prototype.

right and defaults to 10 seconds. When it expires, the low-power sleep mode is activated. Normal operation is resumed by pressing any button.

Screen 2 shows the RES mode, with a 510Ω resistor connected. The three dashes at upper right indicate that the timer is paused. That means the ESR Test Tweezers will not go to sleep; it will probably drain the battery within a day or two if left like this.

Screen 3 shows a 10μF capacitor connected in CAP mode; similarly, the timer has been paused to allow continuous readings to be made. All three operating modes also show the cell voltage at the top of the screen.

Our prototype could function down to around 2V. This is about the point at which the PIC24 processor stops working. We specify 2.4V as the minimum supply voltage, as the accuracy of readings declines significantly below that.

A long press of S3 (about two seconds) switches between operating and settings modes, with S3 then cycling through the various parameters and S1 and S2 adjusting them. The ESR Test Tweezers are usable without calibration, but the calibration steps are easy. There are also a couple of customisation preferences you can apply.

Many calibration steps involve measuring a known value or voltage with the Tweezers and trimming the calibration factor until the displayed value is accurate, which is quite simple and intuitive.

The suggested parts to use are 10Ω, 100Ω and 1kΩ resistors for calibrating ESR and a 10μF capacitor for calibrating capacitance. These values are near the top of their ranges, so they will provide the best resolution when performing the calibration.

The calibration factors are shown in ohms because they are analogous to providing an exact value for the second resistor in a divider. However, because of the circuit's complexity, they don't correspond to any measurable resistance value.

If you don't have these exact value resistors, a lower value (preferably within that decade) will be adequate. Higher values might be outside the limit of their respective range, in which case the display will show "OPEN".

UP
<S1
DOWN
<S2
MODE
<S3
HOLD FOR SETTINGS
SILICON CHIP
ESR TWEEZERS
©2024
04105242
Tweezers Panel

We designed this PCB to protect the back of the Test Tweezers. It can be attached to the cell holder with double-sided tape. It has markings on the opposite side so that it can also be used for the Advanced Test Tweezers. This blue version will be available on our website for users of the Advanced Test Tweezers, although a white version will be included in ESR Test Tweezers kits.

Remember that while resistors are readily available with 1% tolerance or better, capacitors could vary up to 20%. If possible, measure your capacitor with an accurate capacitance meter and use that instead of the nominal value.

The panels above with Screens 4-12 detail the available calibration and setup options. Be sure to do the steps in the order listed, as some factors depend on others being set accurately beforehand.

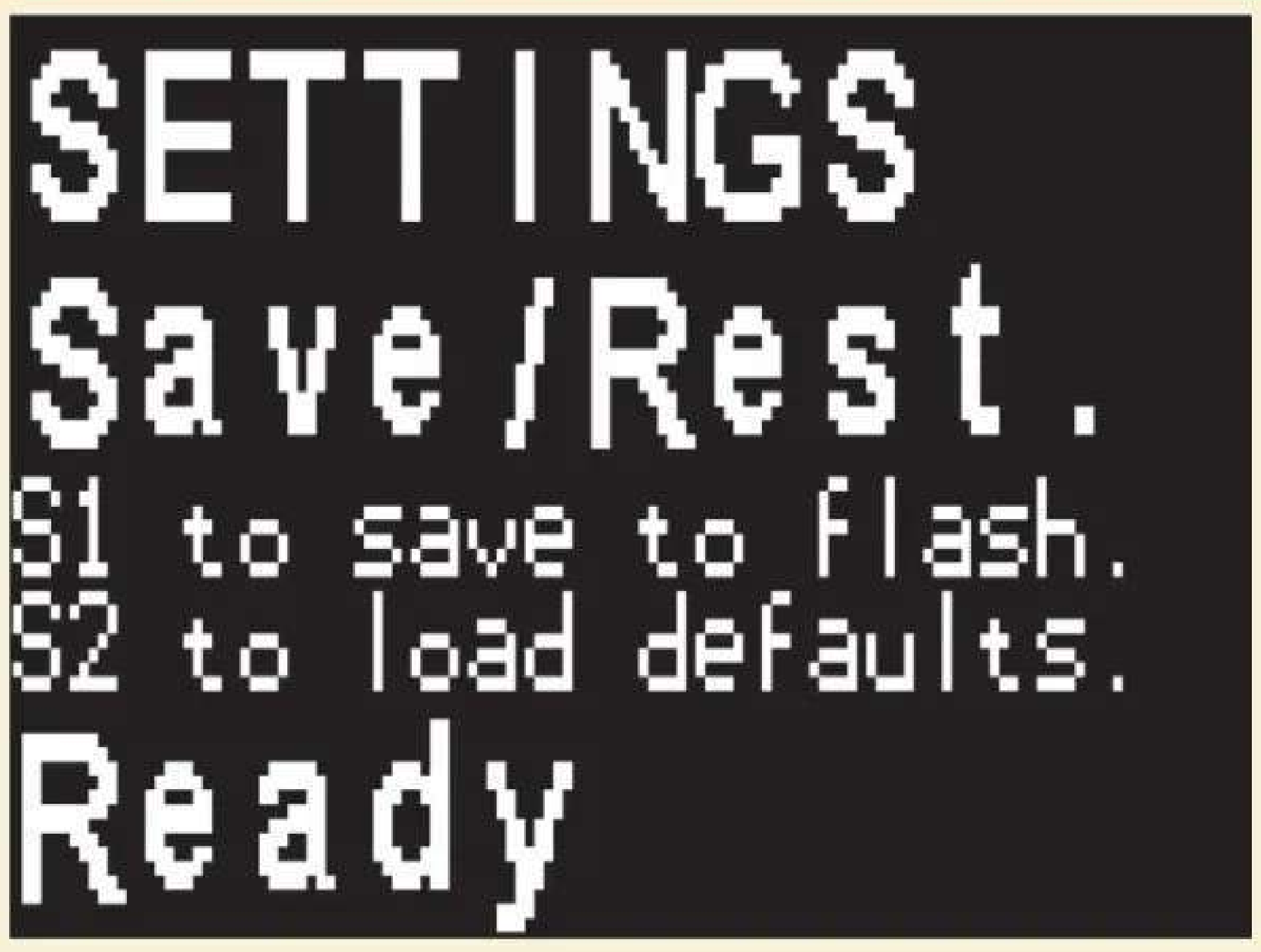
To return to normal operation from settings, press and hold S3 for about two seconds. Be aware that the sleep timer does not count down while in Settings mode, so you should return to operating mode immediately after changing the settings to avoid draining the battery.

Using the ESR Tweezers

Connect the component to be tested between the tips of the probes and apply pressure to make sure they are making good contact. Polarised components should have their positive lead connected to the top (TP+) tip. However, the test voltage is low and should not cause damage if the component is reversed.

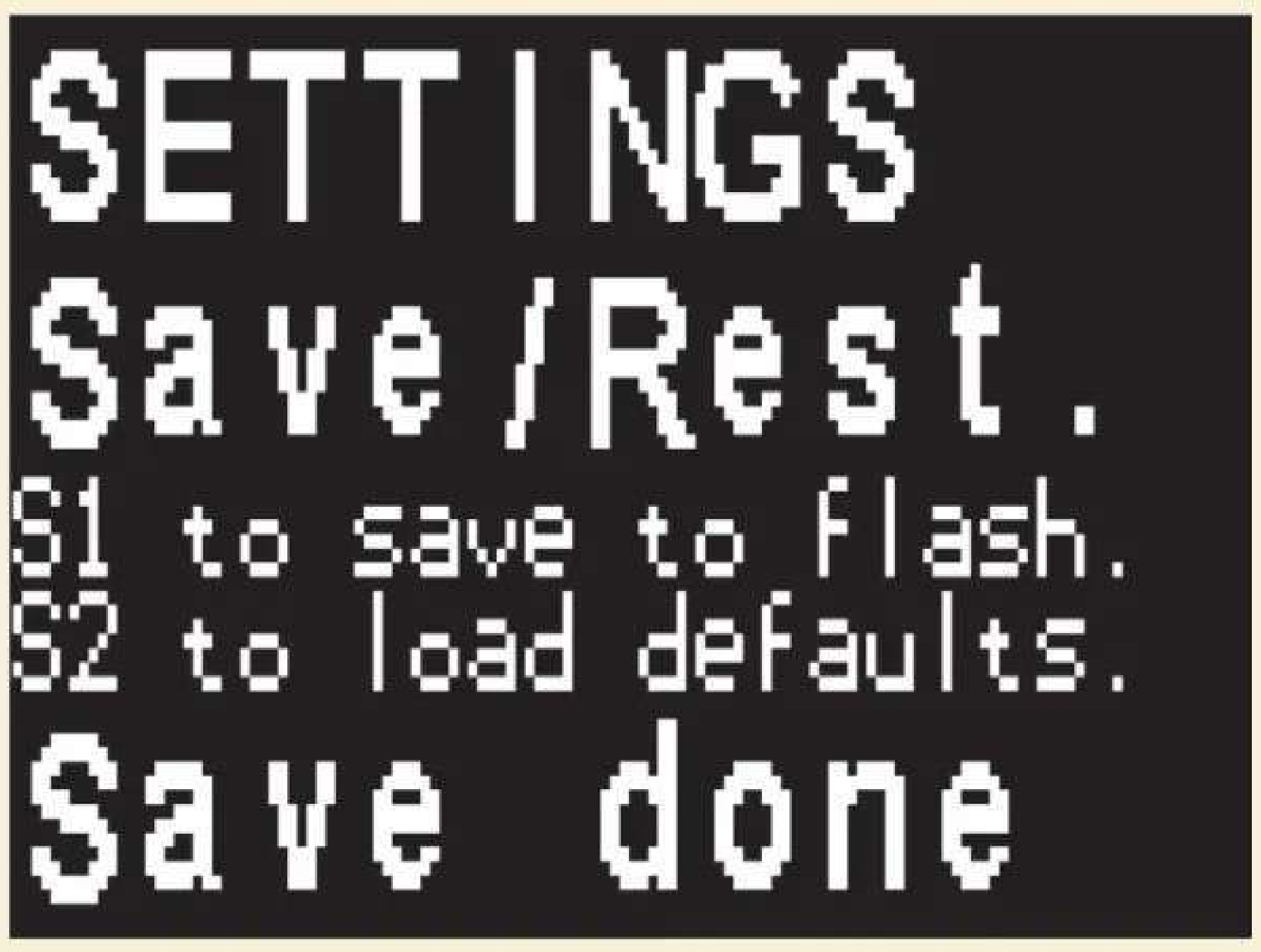
Diagnosing capacitor problems due to high ESR is helpful for those in the power and audio fields. Now you can check that with a handy, compact tool that doesn't cost much to build.

The ESR Test Tweezers can measure ESR, resistance & capacitance (albeit over somewhat limited ranges), making them more valuable than the 2004 design and in a smaller package. SC



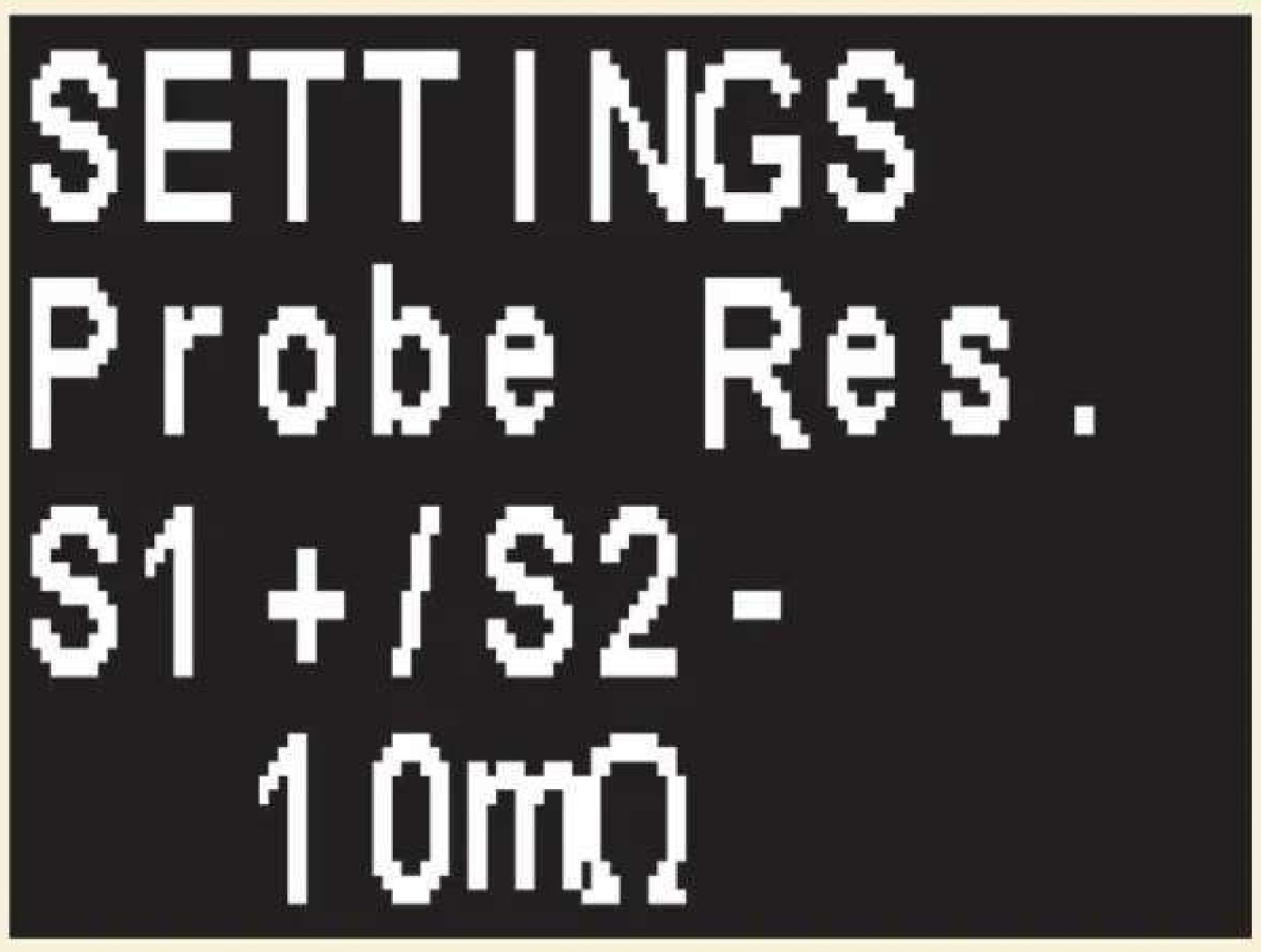
SETTINGS
Save/Rest.
S1 to save to flash.
S2 to load defaults.
Ready

Screen 10: this is the last screen you should need to use for setup and calibration. Press S1 to save any altered settings to flash memory; S2 will load the defaults in case the saved data becomes corrupted. The defaults can also be loaded by holding S3 while powering up the Tweezers.



SETTINGS
Save/Rest.
S1 to save to flash.
S2 to load defaults.
Save done

Screen 11: after saving to or restoring from flash, you should get a message indicating it completed successfully. This is the last necessary step for setup and calibration; a long press of S3 will return to operating mode. As well as on the first use, you should recalibrate when a new cell is fitted.



SETTINGS
Probe Res.
S1+/S2-
10mΩ

Screen 12: there are some screens after Save/Restore that should not need to be changed; they adjust the factors set by the Calibrate step shown in Screen 4. They include the probe contact resistance (shown here) and two pages with ADC limit values, used to check that readings are valid.

Dissipation factor, loss angle and ESR

Dissipation factor (DF) and loss angle (δ) measure the energy lost in an oscillating system. Many capacitor data sheets specify these instead of providing an ESR value.

In our case, the dissipation factor and loss angle specifically refer to the losses in a capacitor due to ESR. These terms are also used in other contexts in electrical engineering, but we are looking specifically at capacitor ESR. We want to relate the capacitive reactance to the pure resistance due to ESR. Both can be plotted on the complex number plane, hence the references to angles.

The loss angle is simply the inverse tangent function of the dissipation factor; thus, you might also see 'tangent of loss angle', which means the same as 'dissipation factor'.

Since the reactance changes with frequency, we need to focus on a specific frequency.

For example, in a transformer-based mains power supply, the capacitors will be subjected to predominantly 100Hz (50Hz mains) or 120Hz (60Hz mains) ripple. Capacitors in audio circuits will be subjected to a broader range of frequencies, perhaps 20Hz to 20kHz. Capacitors in switch-mode supplies will generally have ripple at 20kHz to 2MHz.

Let's take a concrete example of a capacitor, such as the 4700μF 50V electrolytics we have previously in our projects, in the form of Nichicon UVZ1H472MRD capacitors. We used them to filter the rectified output of a mains transformer. Their data sheet lists a (maximum) tangent of loss angle of 0.2.

That corresponds to a loss angle of 11.3° or 0.197 radians, ie, tan(11.3°) ≈ 0.2. Note that the loss angle (in radians) is very close to the dissipation factor for typical values. This is a well-known approximation for the tangent function at low values.

Using the impedance equation for capacitors of $Z = 1 \div (2\pi fC)$, we get an impedance value of 0.34Ω for a 4700μF capacitor at 100Hz. Multiplying this by the dissipation factor of 0.2 gives an ESR of 0.068Ω, close to the 0.05Ω noted in Table 1 for similar capacitors. If you measured an ESR of 0.05Ω for such a capacitor, that would be acceptable, as it is below the specified maximum.

The loss angle (δ) can be visualised with a diagram of the complex impedance (Fig.a), which shows the reactance due to capacitance in the imaginary plane (vertical) and the resistance due to ESR in the real plane (horizontal).

The cosine of the loss angle relates to the proportion of energy transmitted by the capacitor (compared to that dissipated by the ESR). At low loss angles, the cosine of δ is close to unity, and there are no losses, although they rise sharply as the angle (and ESR) increases.

These ideas are similar to concepts like power factor (and power angle), although, in AC power systems, the capacitive element is undesirable and a purely resistive load is preferred.

You can also see from this how a high ESR would create a phase shift for audio signals, increasing distortion.

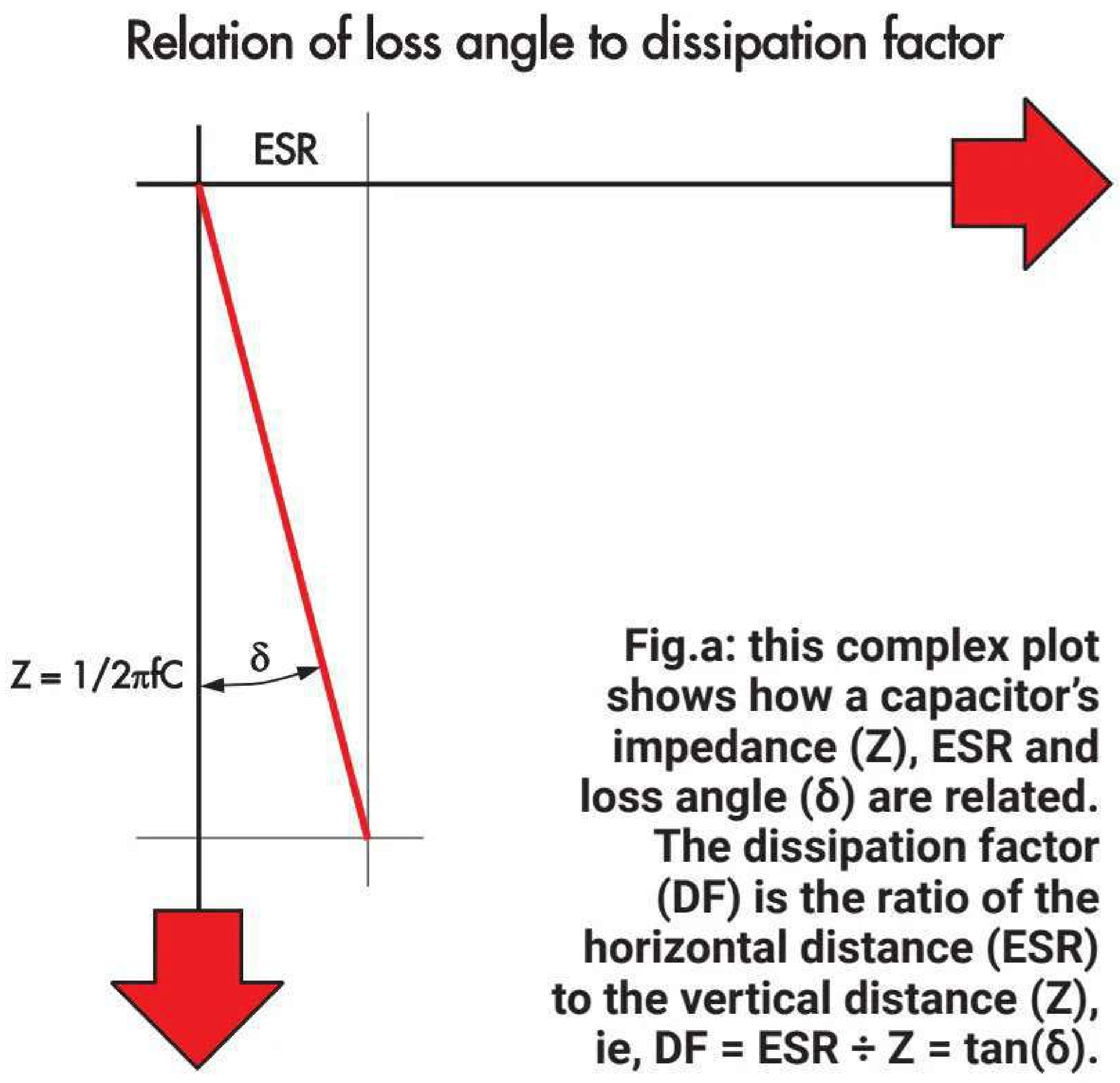


Fig.a: this complex plot shows how a capacitor's impedance (Z), ESR and loss angle (δ) are related. The dissipation factor (DF) is the ratio of the horizontal distance (ESR) to the vertical distance (Z), ie, DF = ESR ÷ Z = tan(δ).

AUDIO OUT

By Jake Rothman

Using Capacitors for HiFi, Part 2

This series started last month by explaining how suitable different types of capacitor are for audio use. In this second and final instalment, we'll continue that investigation, concentrating on some of the more unusual types of capacitors for HiFi, and give a ranking of all the major types.

Non-polarised solid tantalum capacitors exist, but don't be fooled, they are just two standard tantalum capacitors in one case (see Fig.24). I sometimes solder a bias wire to the case.

I have also seen capacitors wired in inverse parallel. Of course, this does not work well with voltages above 1V either way, but it can be useful for blocking very low offset voltages while maintaining a high capacitance. The distortion performance is shown in Fig.26; it's about half that of a single capacitor of the same type.

When using back-to-back polarised capacitors, you might expect the capacitance of the pair to be half that of an individual unit, but that is not necessarily so. Depending on the biasing and signal level, one capacitor may be acting like a diode at any time if it is sufficiently reverse-biased. Switching in and out of that mode can introduce extra distortion.

This is mainly a problem with series back-to-back polarised capacitors in long time-constant integrators, such as LFOs in sweep generators. One way to avoid this is to provide a centre-bias to the junction of the capacitors. With audio coupling, where there is little voltage drop across the pair, the capacitors will act more like you would expect.

The general consensus appears to be if the frequency is above 1Hz and the voltage across the composite bipolar capacitor is kept below about 10% of the rated voltage, there is no capacitance doubling.

Bypassing

It's often said by HiFi buffs that bypassing an electrolytic capacitor with

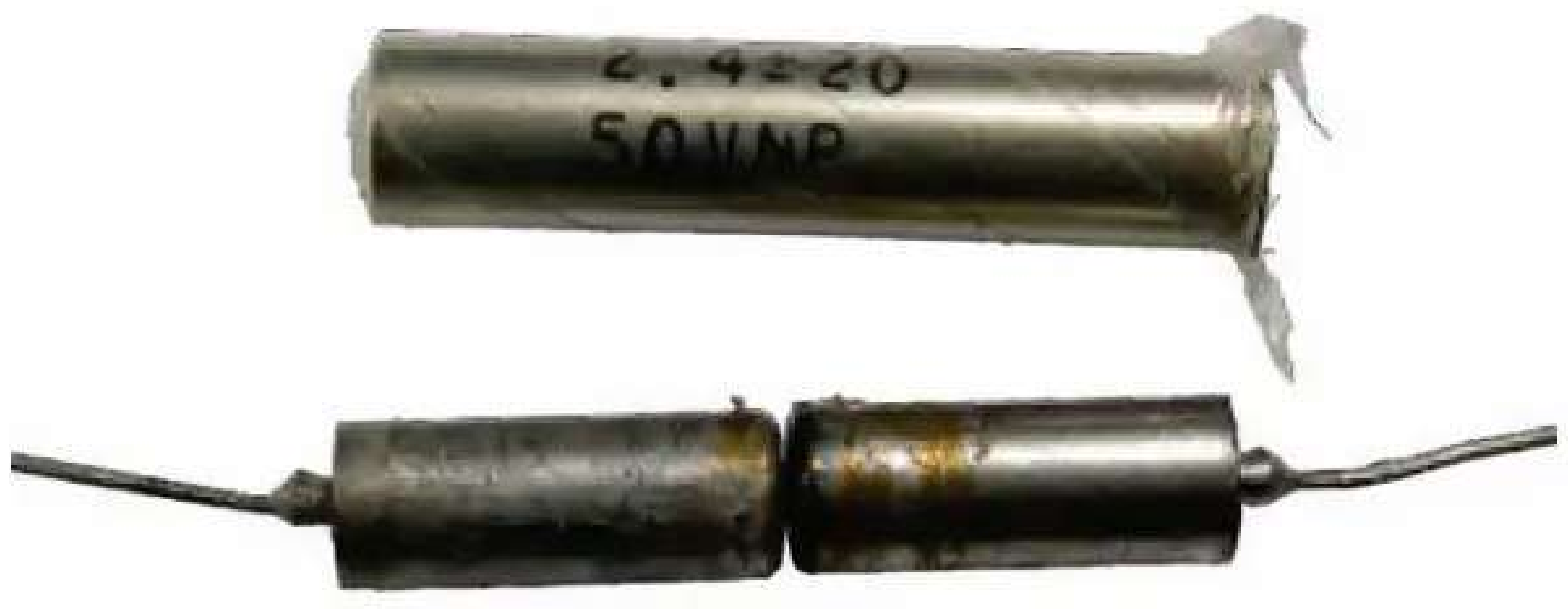


Fig.24: a bipolar tantalum capacitor. I was shocked to find it was just two standard polarised units in one case!

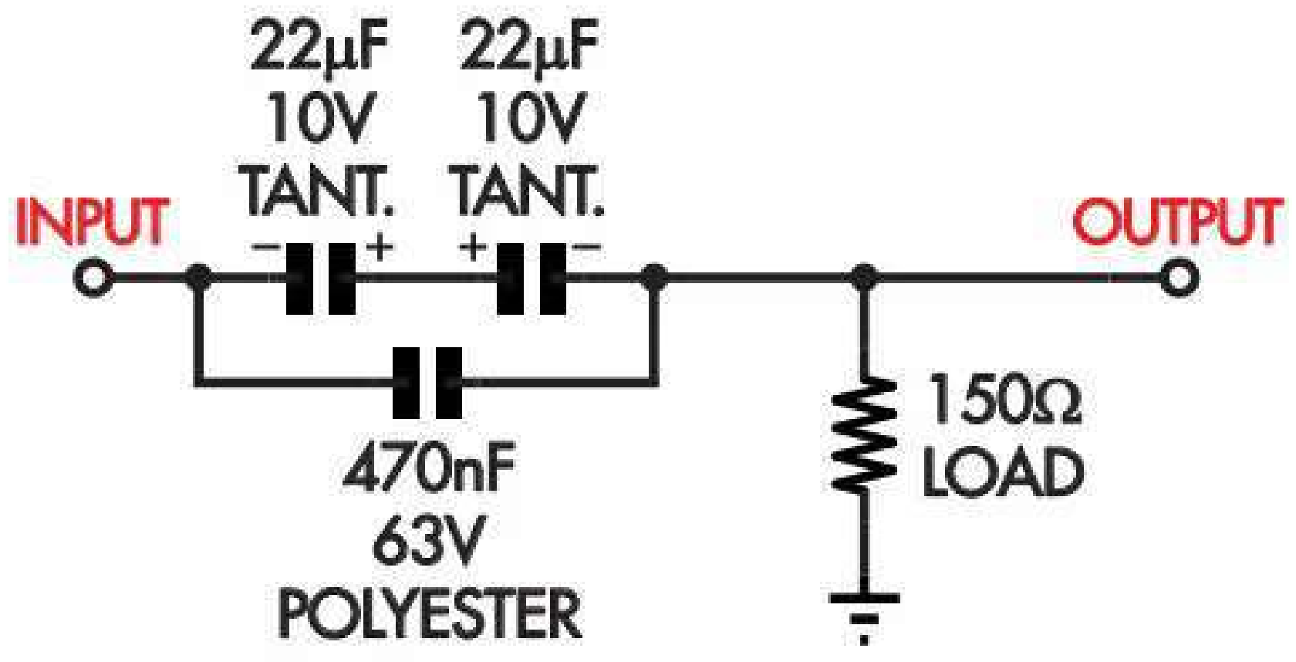

22µF
10V
TANT.
22µF
10V
TANT.
INPUT
OUTPUT
470nF
63V
POLYESTER
150Ω
LOAD

Fig.25: bypassing two back-to-back tantalums with a 470nF polyester cap.

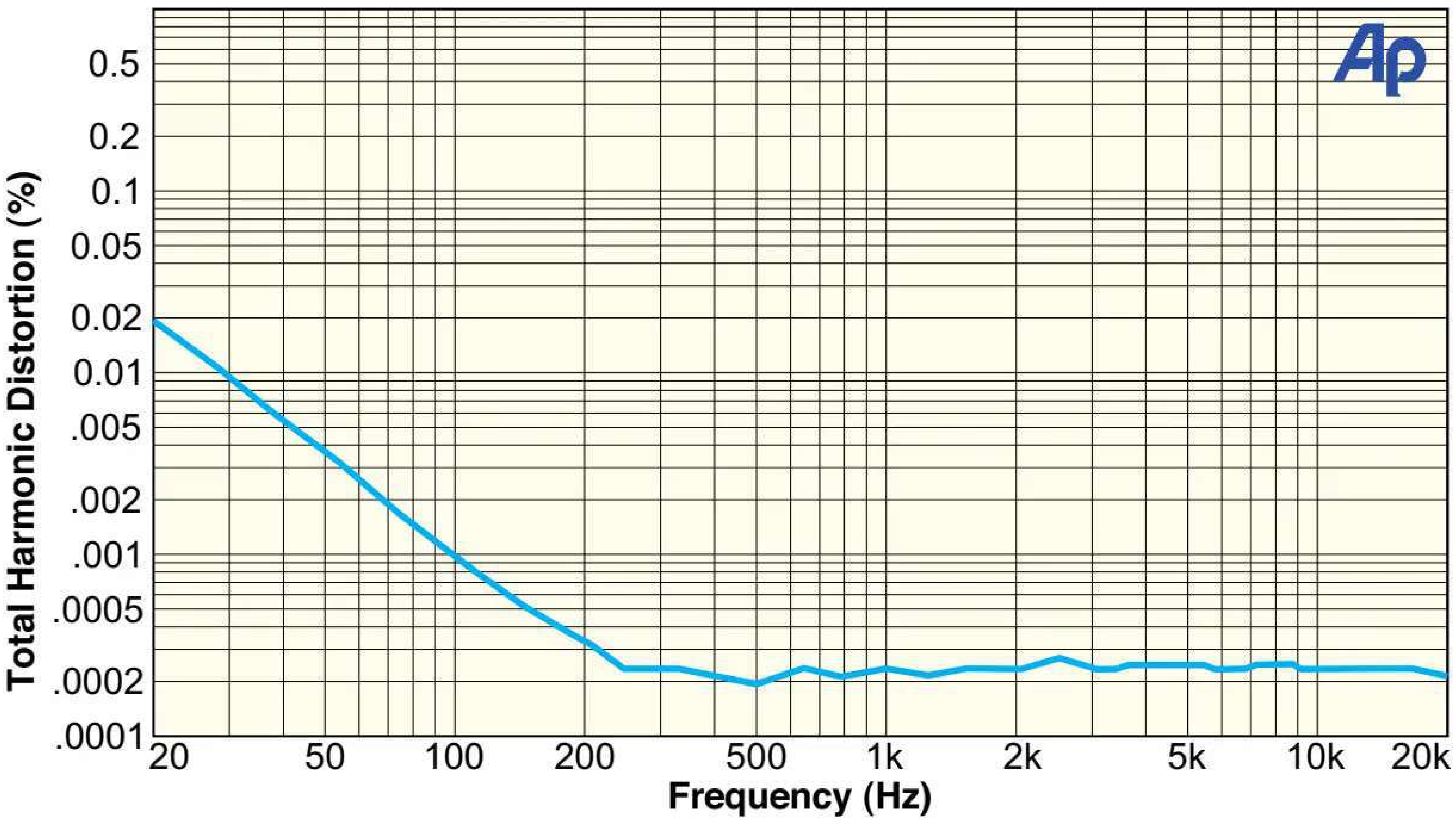


Fig.26: the distortion curve of a reverse parallel network (two 47µF 10V caps). It isn't as bad as expected, seeing as they were being thrashed at 6V RMS.

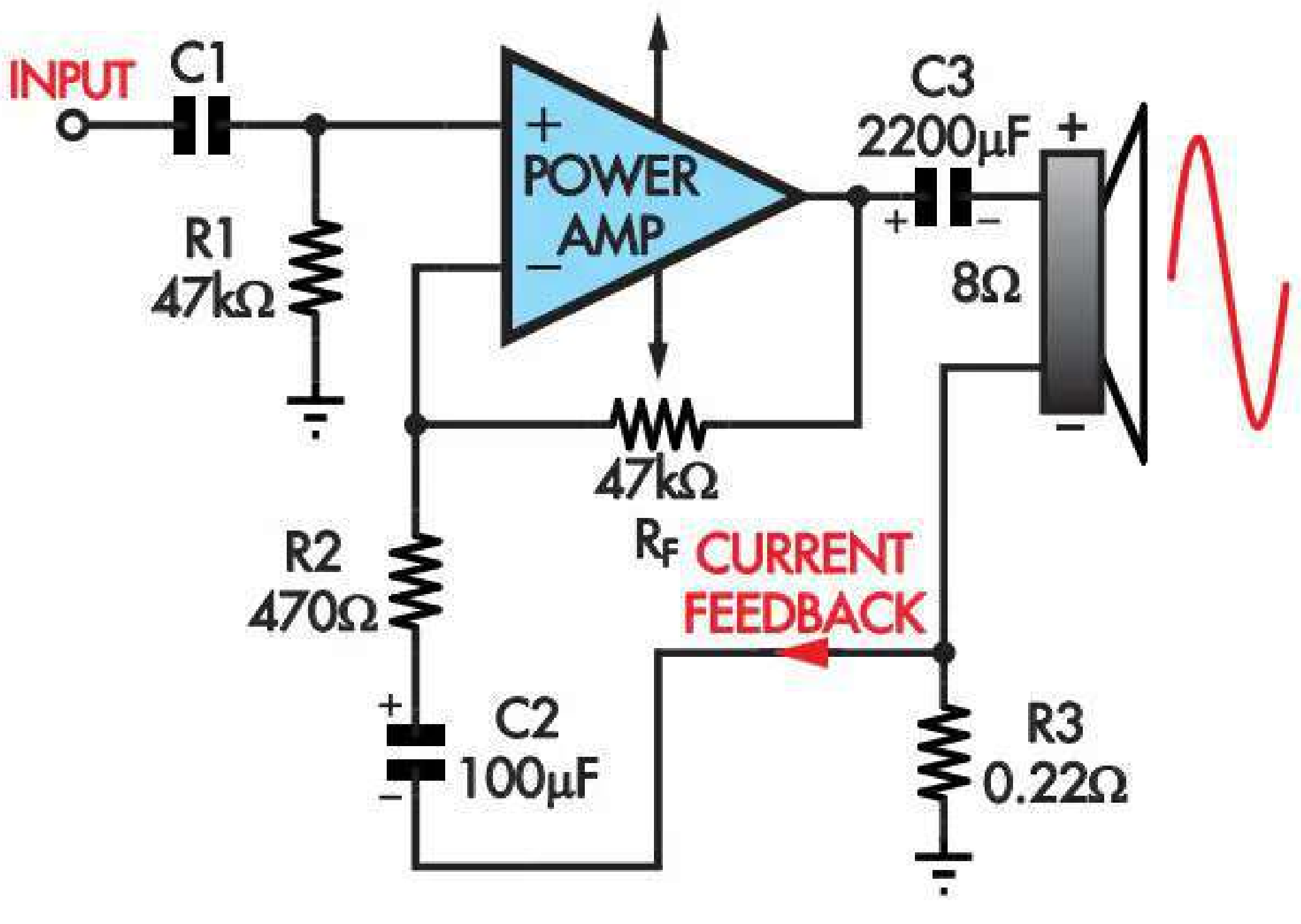

INPUT
C1
R1
47kΩ
POWER
AMP
C3
2200µF
8Ω
47kΩ
R_F
CURRENT
FEEDBACK
R2
470Ω
C2
100µF
R3
0.22Ω

Fig.27: this power amplifier circuit applies feedback around the output electrolytic, linearising it.

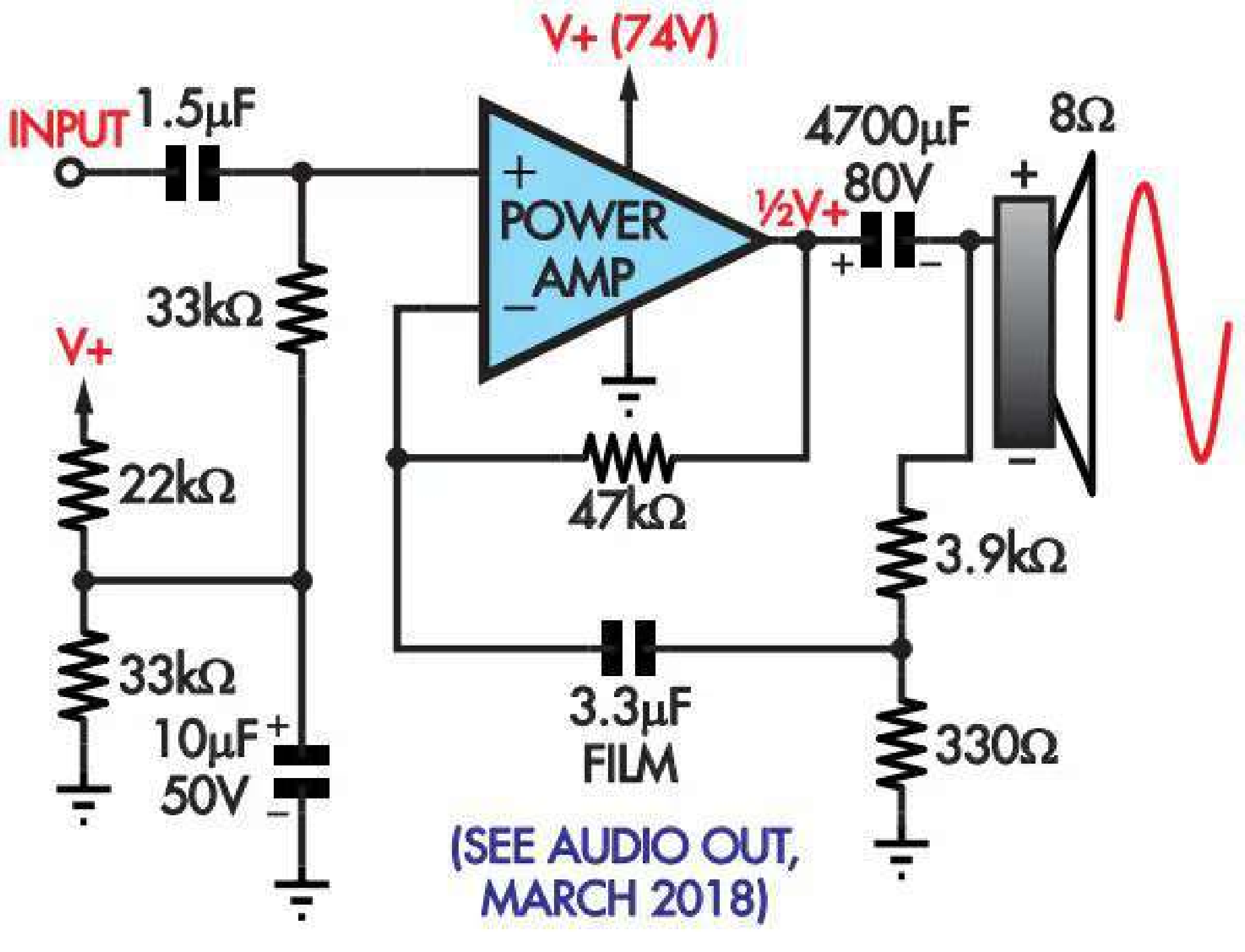

V+ (74V)
INPUT
1.5µF
33kΩ
V+
22kΩ
33kΩ
10µF
50V
POWER
AMP
4700µF
80V
½V+
8Ω
47kΩ
3.9kΩ
3.3µF
FILM
330Ω
(SEE AUDIO OUT,
MARCH 2018)

Fig.28: low distortion can be obtained using a standard 4700µF 80V output capacitor in the amp's feedback loop.

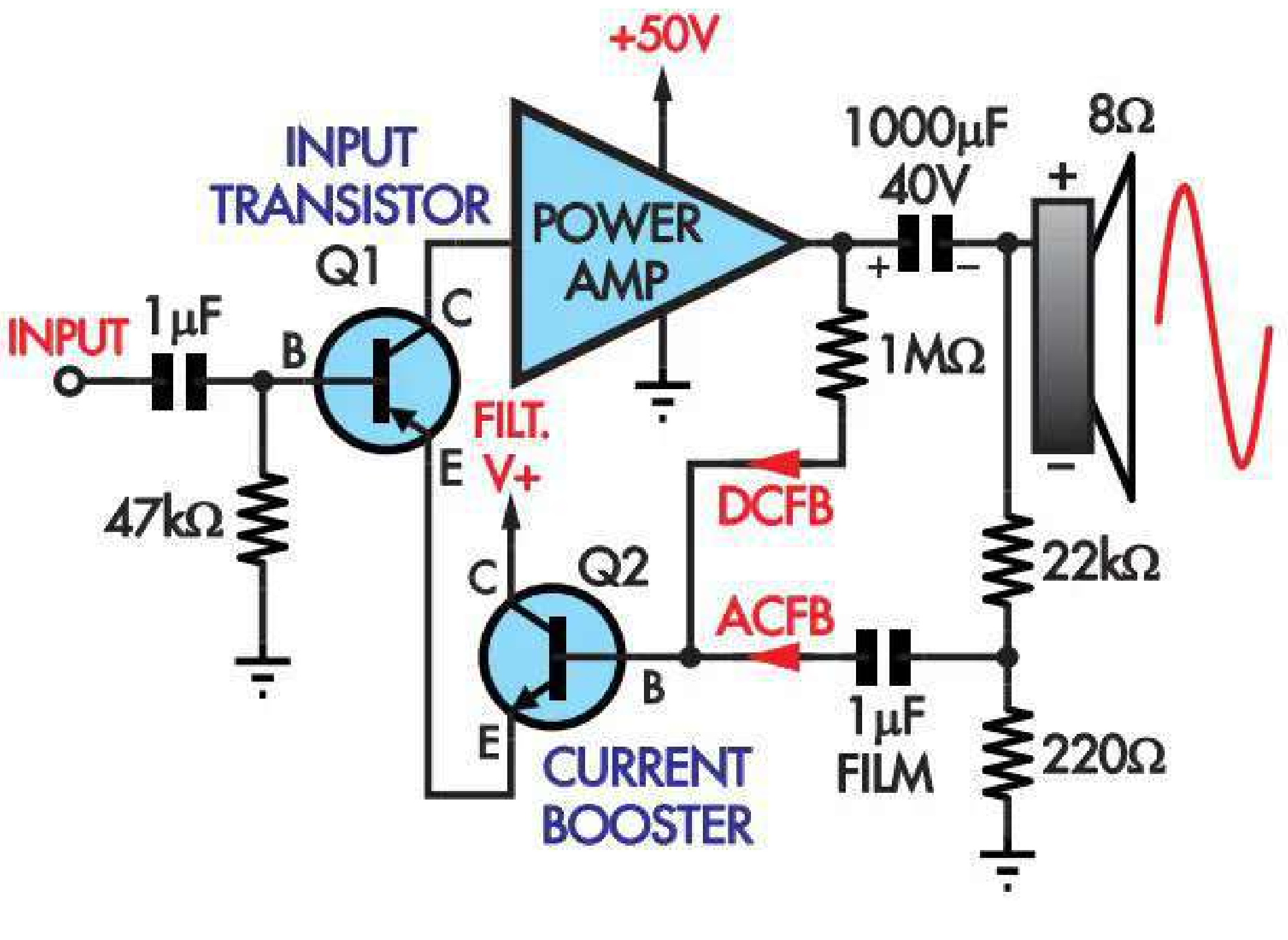

+50V
INPUT
TRANSISTOR
Q1
POWER
AMP
1000µF
40V
8Ω
INPUT
1µF
B
C
E
47kΩ
1MΩ
FILT.
V+
DCFB
Q2
C
B
E
ACFB
1µF
FILM
22kΩ
220Ω
CURRENT
BOOSTER

Fig.29: G. Nalty's solution for linearising the output capacitor (Q2 allows the DC feedback resistor value to be high).

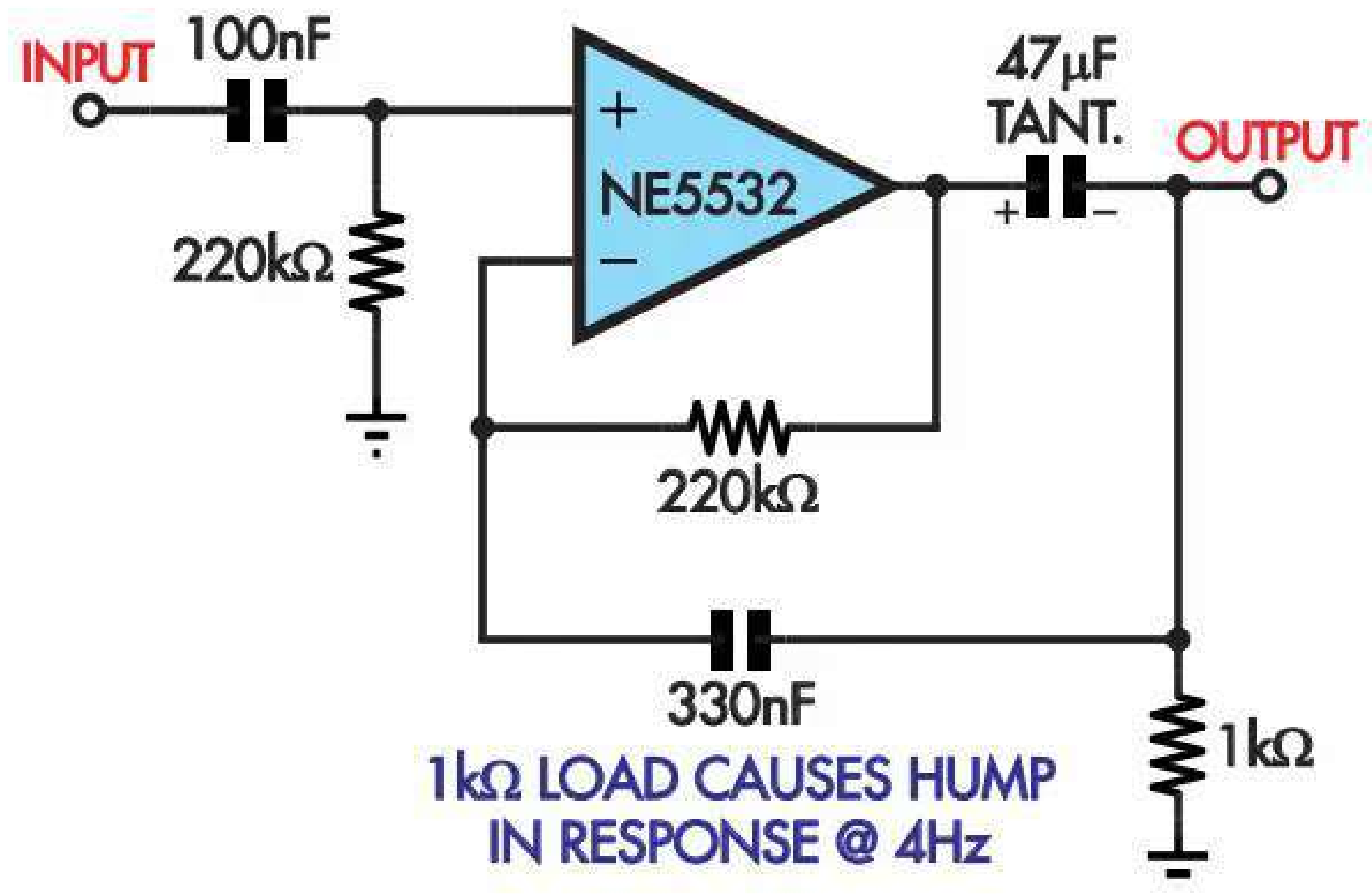


Fig.30: an op amp follower with its electrolytic output coupling capacitor in the feedback loop.

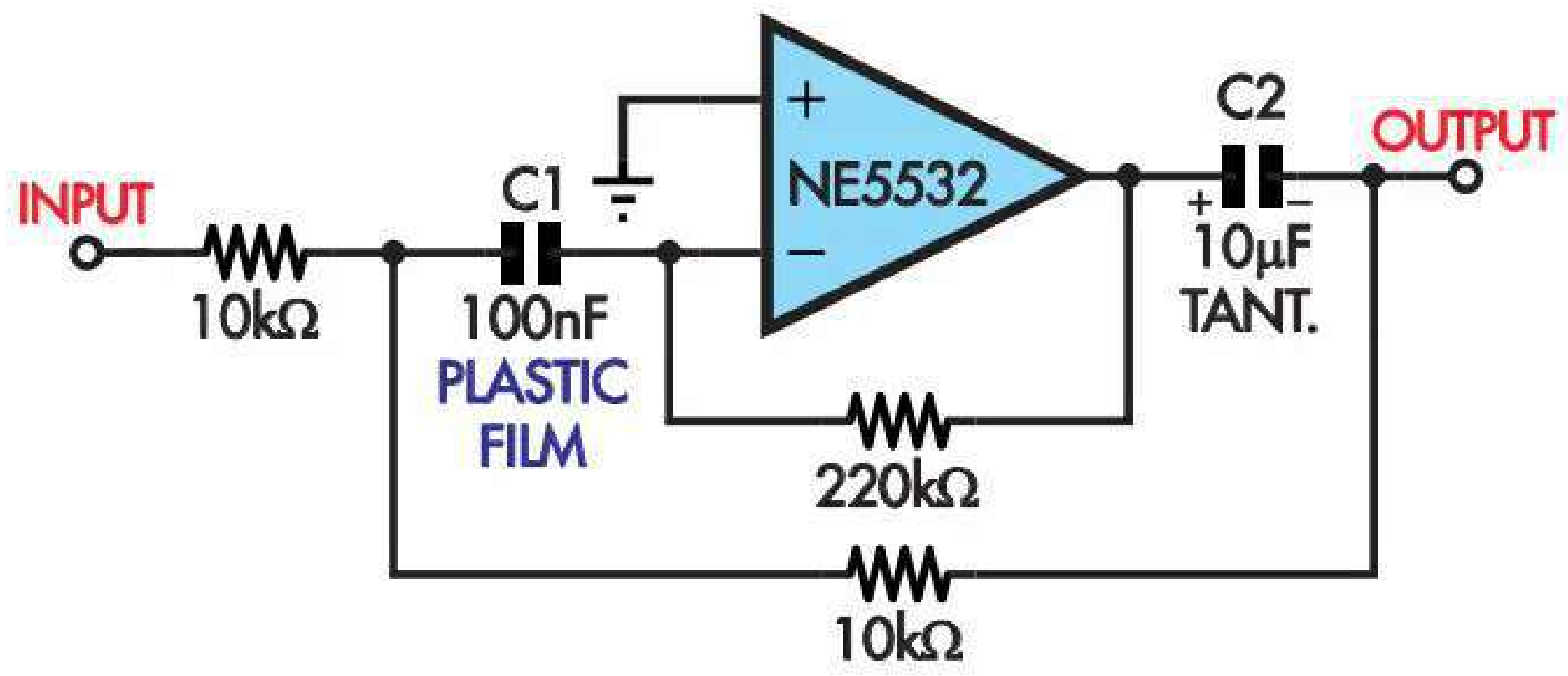


Fig.31: an inverting amplifier trick to bootstrap C1 on a virtual earth to increase its effective value by 10 times,

a plastic film type can reduce distortion, as shown in Fig.25. Since capacitor distortion is a low-frequency effect, it doesn't really work. It does work for high-frequency decoupling, however.

When I made some measurements, it did reduce some of the high-order distortion harmonics when gross LF distortion was occurring by about 30%. But if that's happening in your circuit, it probably isn't what you'd call HiFi, anyway.

Negative feedback

Capacitor distortions produced within a negative feedback (NFB) loop, such as driver-stage bootstrap capacitors, emitter bypass, and the V_{BE} multiplier bypass in power amps are all reduced by the feedback. Feedback compensation capacitors, such as C_{DOM} (the dominant pole 'compensation' capacitor, usually across the VAS transistor), must be linear, though.

I once used an X7R ceramic capacitor instead of an NP0 type for C_{DOM} and the high-frequency distortion went up by a factor of four. (By the way, NP0 and C0G are different terms for the same specification).

The distortion-reducing powers of NFB can be applied to output capacitors on power amplifiers. The distortion reduction is usually in the order of 10 times (20dB). I discovered this effect when making constant-current power amplifiers, as shown in Fig.27. Including the output electrolytic in the NFB loop reduced the distortion.

NFB around the output capacitor works well with normal power amps, as shown in Fig.28.

Fig.29 shows a technique developed by engineer/*PE* writer Graham Nalty for single-rail amplifiers with a single input transistor. He used this in his Class One Sound amplifier in *Everyday Electronics*, January 1989. Op-amp followers (Fig.30) can also have the output capacitor included in the loop.

Another trick is to apply NFB to input capacitors on inverting amplifiers, as shown in Fig.31. This allows a smaller capacitor C1 for a given LF roll-off to be used. Calrec used this in their virtual-earth mixer amplifiers. The output capacitor, C2, can also be included in the feedback loop to linearise it.

The input capacitor is not linearised, but it can be reduced in capacitance enough to be a film capacitor.

Care has to be taken to prevent gain rise at subsonic frequencies when applying NFB around capacitors. The circuit should be tested for LF instability with tone bursts and the required load.

If the capacitor values are not carefully selected so the frequencies are spaced apart, LF bouncing on tone burst and pulses may occur. Sometimes LF oscillation or 'motorboating' can result from the phase shift.

Subtler stuff

While researching his book *The Design of Active Crossovers* (2nd ed, Routledge, 2020), Douglas Self observed that small polyester capacitors improve with use. However, the effect was not permanent. He also showed that the second bootstrapped capacitor (C2) in Sallen-Key active filters was much less sensitive to non-linearity than the first, C1 (Fig.32). This reduces the part cost for high-performance active crossovers.

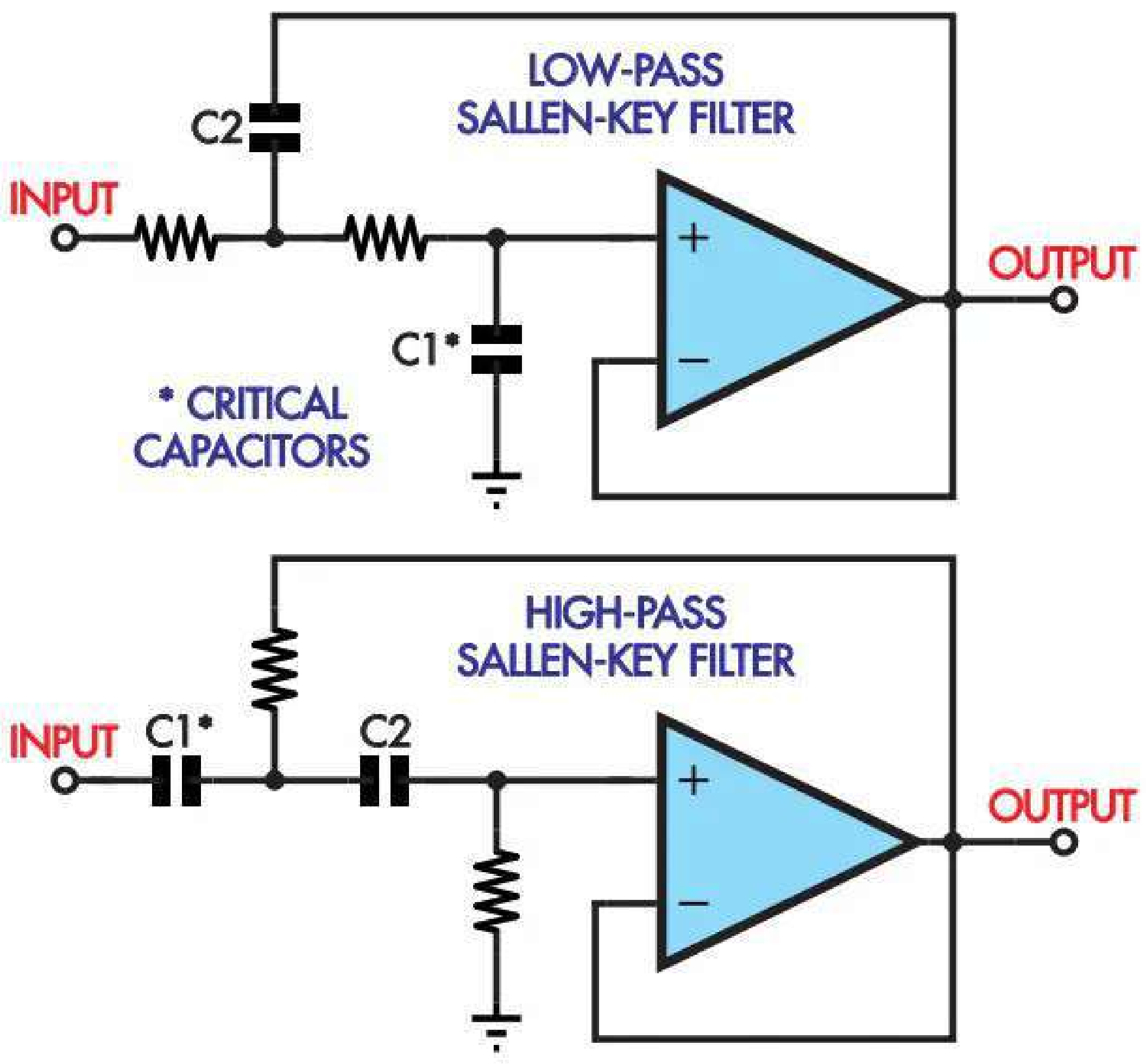


Fig.32: minimising the voltage drop across C2 allows a cheap polyester capacitor to be used in an active filter.

Mechanical construction

The classic axial cylinder construction tends to give the lowest distortion, and most high-quality audio capacitors are of this shape. The late Cyril Bateman, a capacitor engineer at Erie (later ITT) found that flattened oval shapes and stacked-foil forms increased distortion. I suspect this is because the mechanical stresses are less evenly distributed, and there may be regions where movement could occur, leading to self-modulation.

Foil capacitors are large and expensive. They can be inductive if each lead is attached at only one point to the foil. This was later eliminated with the extended-foil technique, where the whole of the foil on the end was soldered together (see Fig.33).

Most plastic film capacitors today are metallised film types, which have higher losses but are smaller and cheaper. They also have the benefit of possessing self-healing properties (Fig.34). Foil capacitors are designated F in the European (DIN) system, with M for metallised. Some Wima foil capacitors are shown in Fig.35.

Capacitor leads

Most leaded components now have copper-clad steel leads. This can

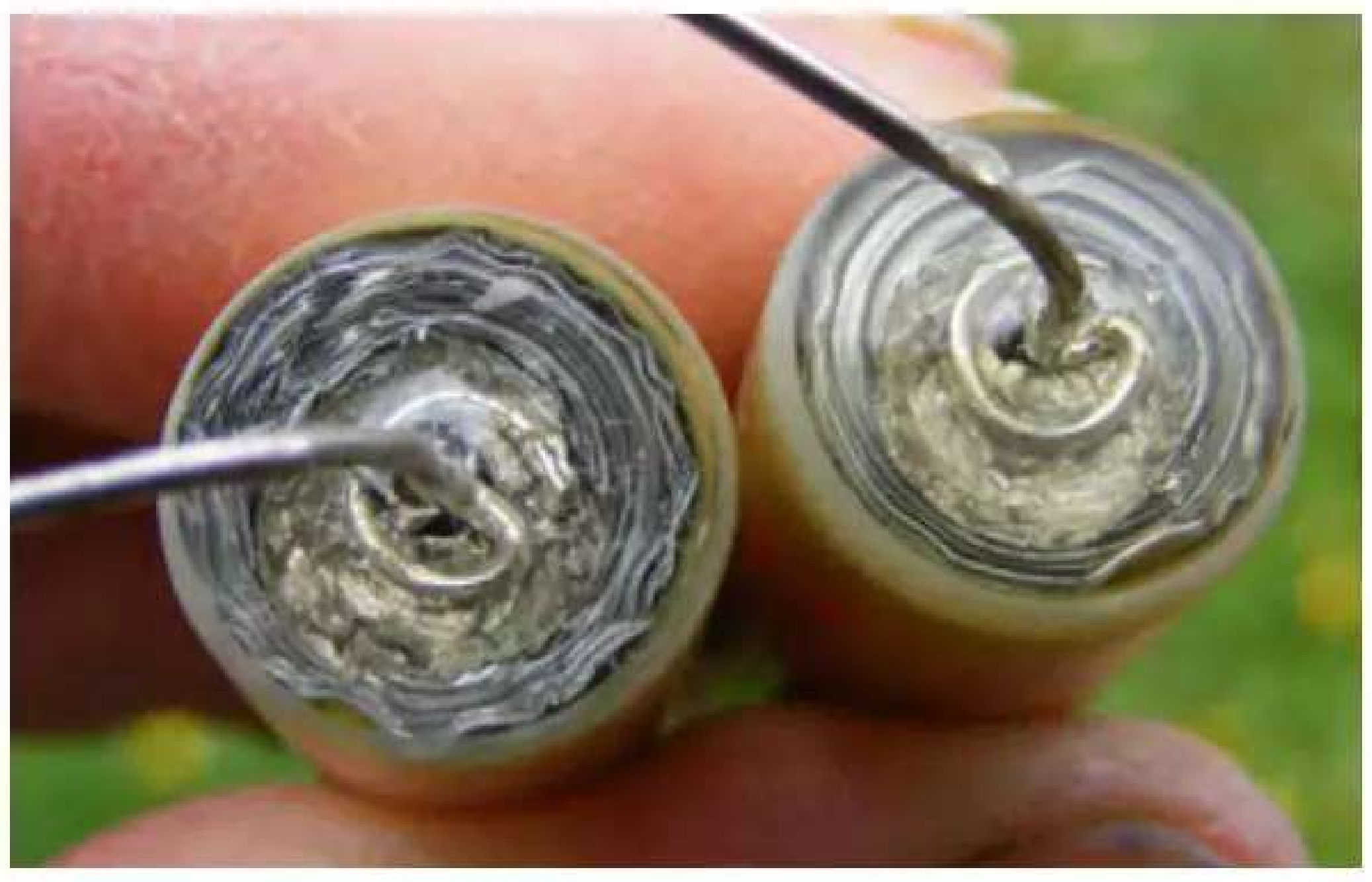


Fig.33: the clear epoxy end-fill on Wima polyester foil capacitors reveals that the contact is on the extended foil.

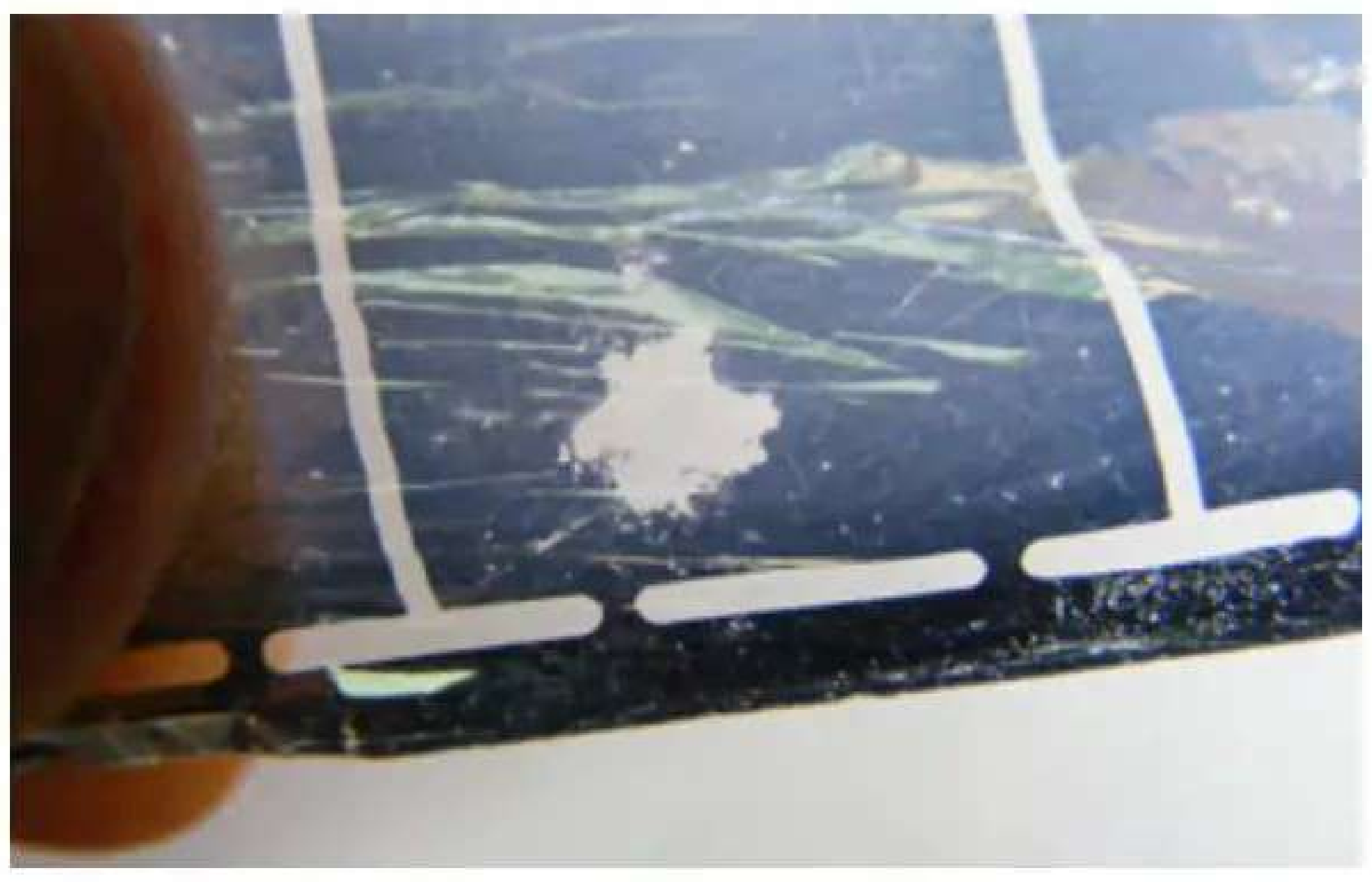


Fig.34: self-healing of a metallised film. Metal burns away from the defect, isolating it (note how thin it is!).

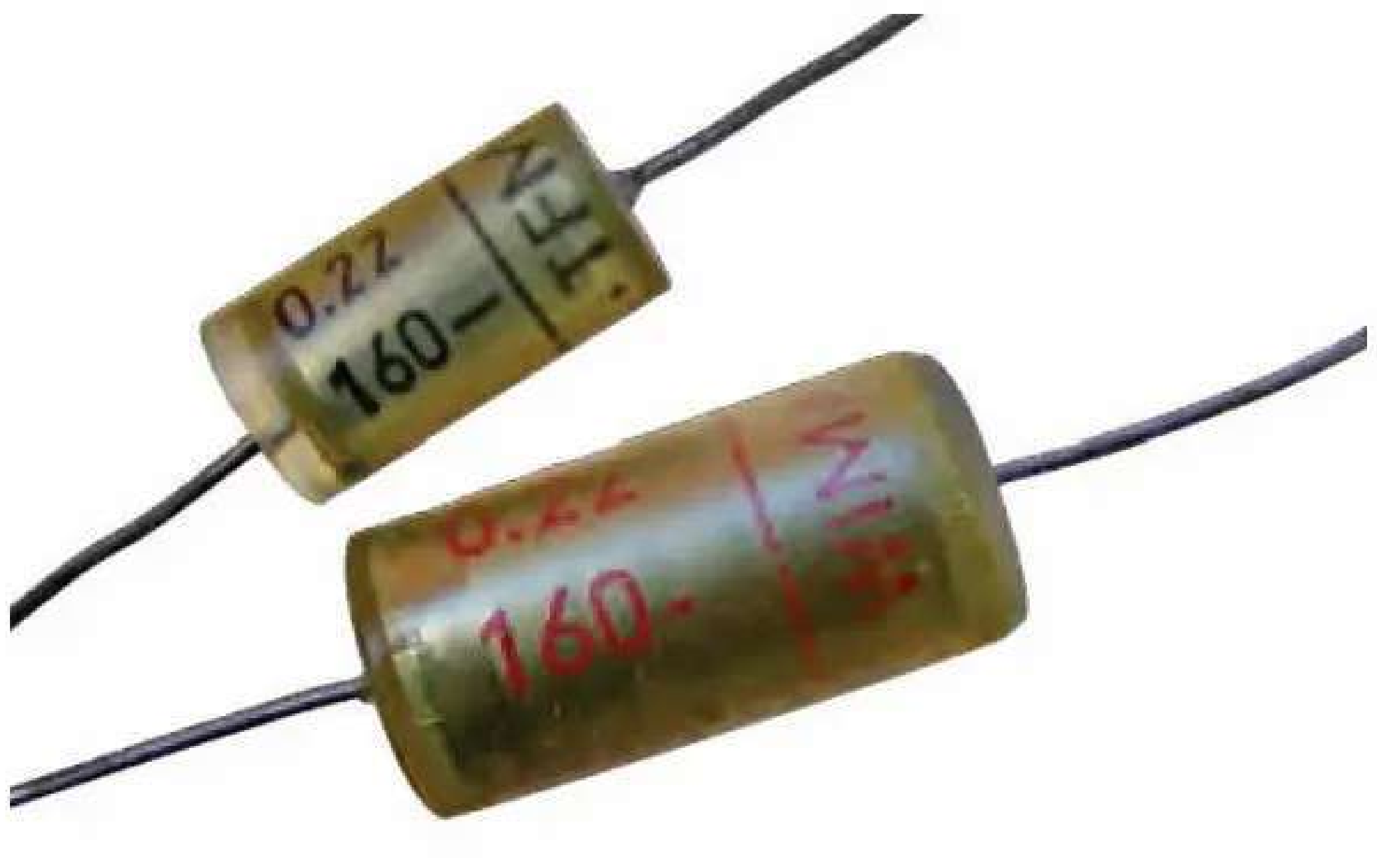


Fig.35: a Wima metal-foil capacitor compared to its smaller metallised version.

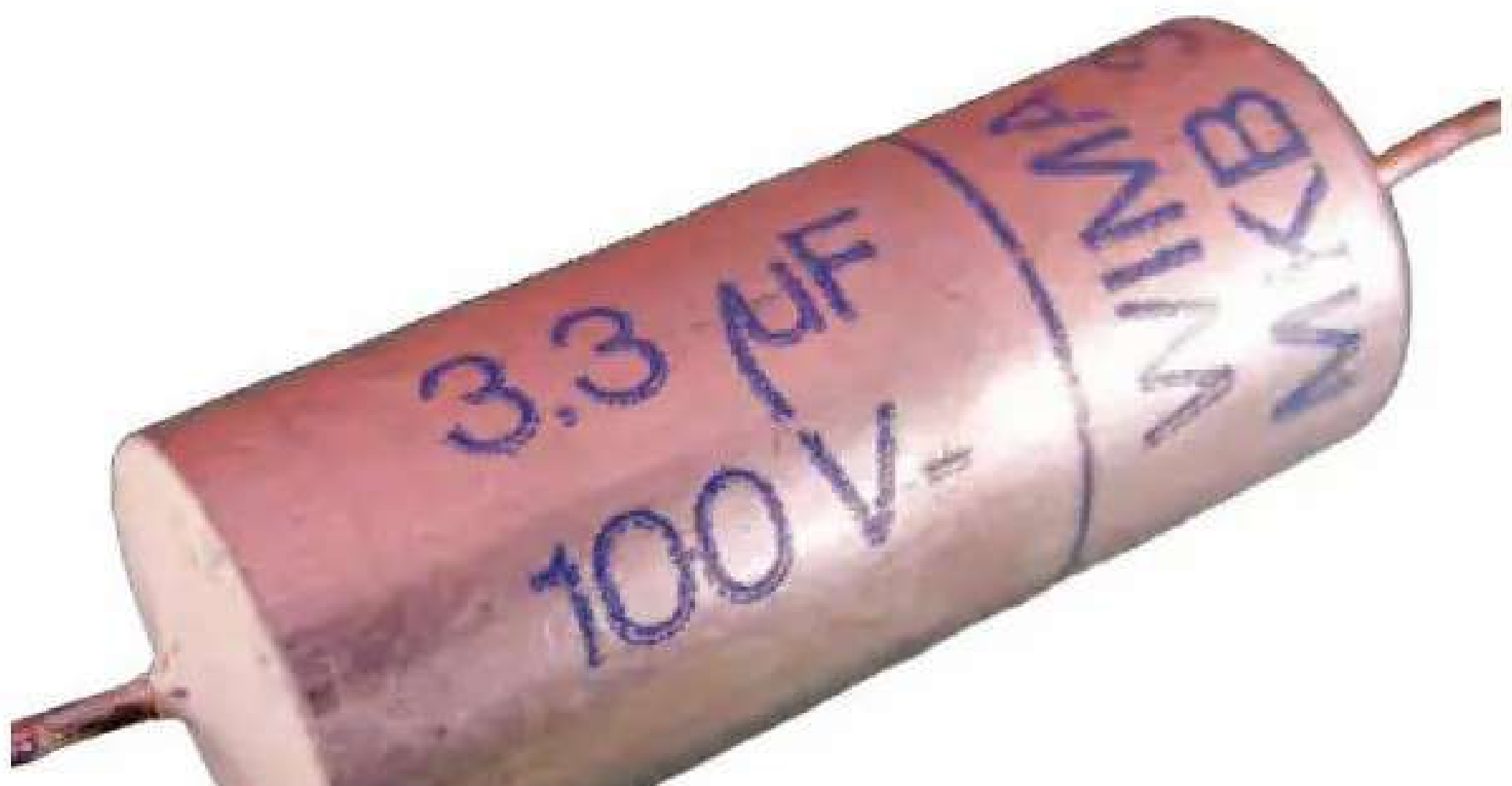


Fig.36: a Wima MKB3 metal-cased polycarbonate capacitor. Unfortunately, mechanical resonance ruined its sound.

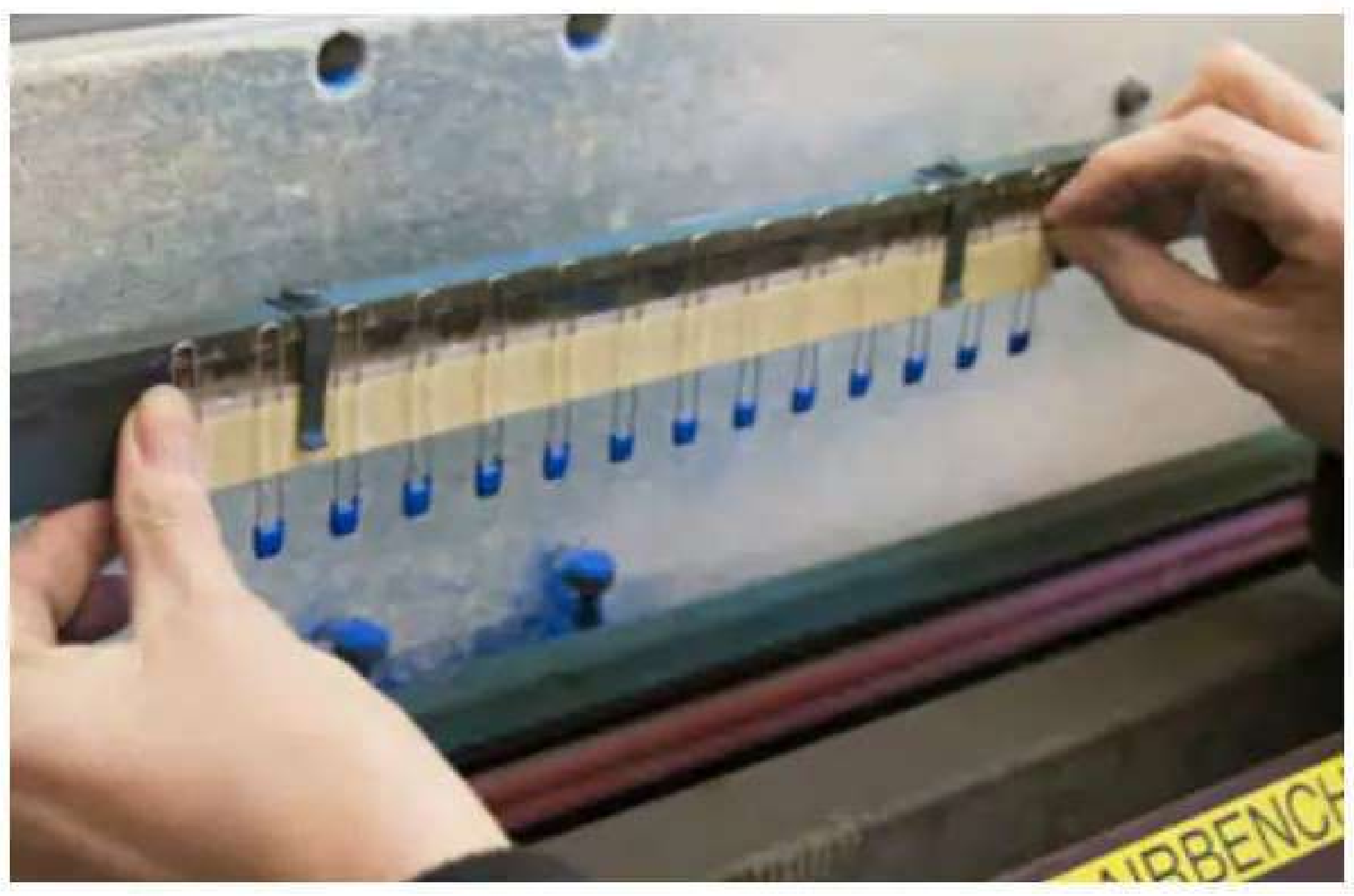


Fig.37: dipping silvered-mica capacitors at Charcroft in epoxy powder. This protects them from moisture etc.

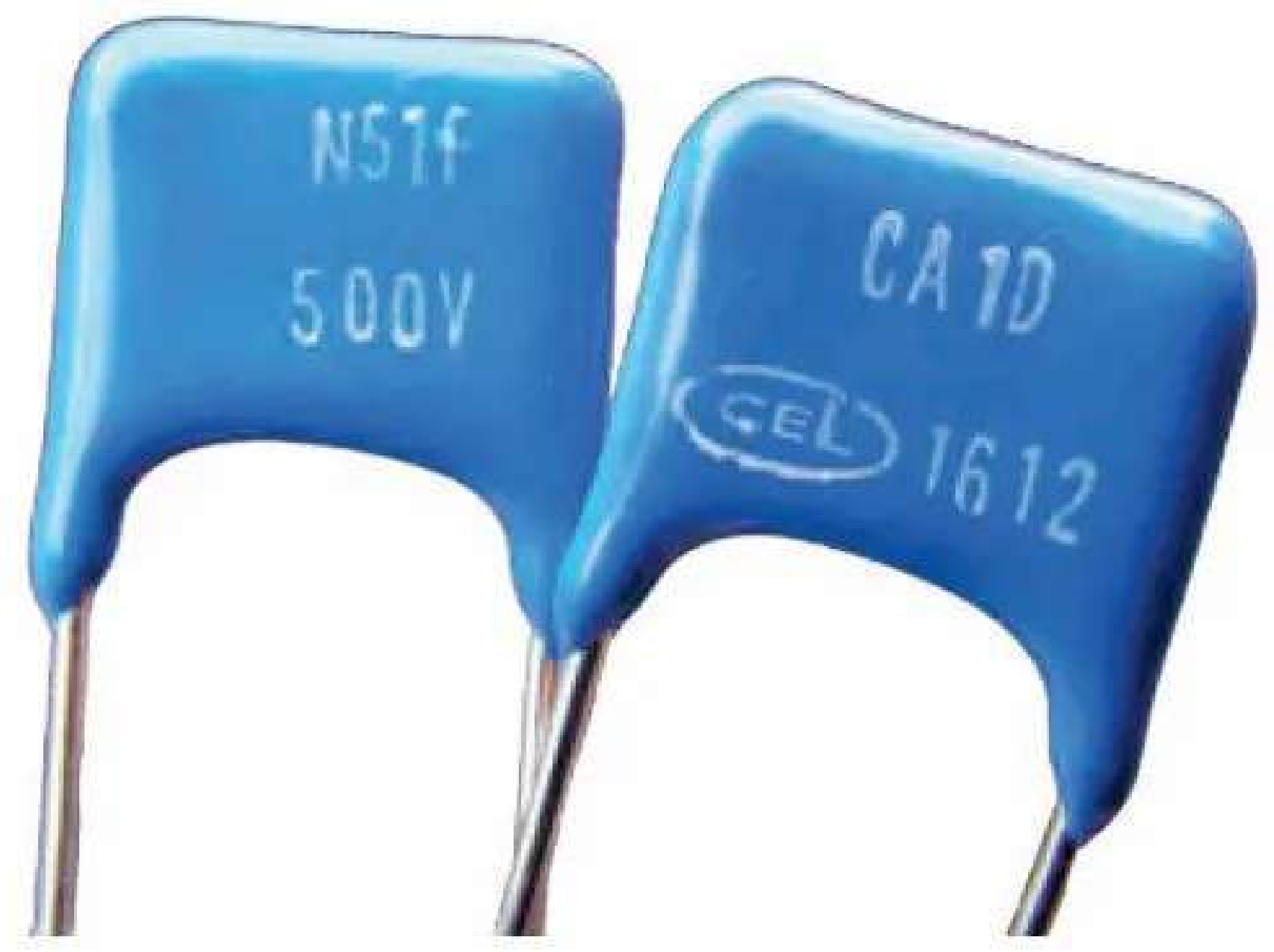


Fig.38: the final encapsulated mica capacitors. Other types of capacitor, like ceramics, are protected similarly.

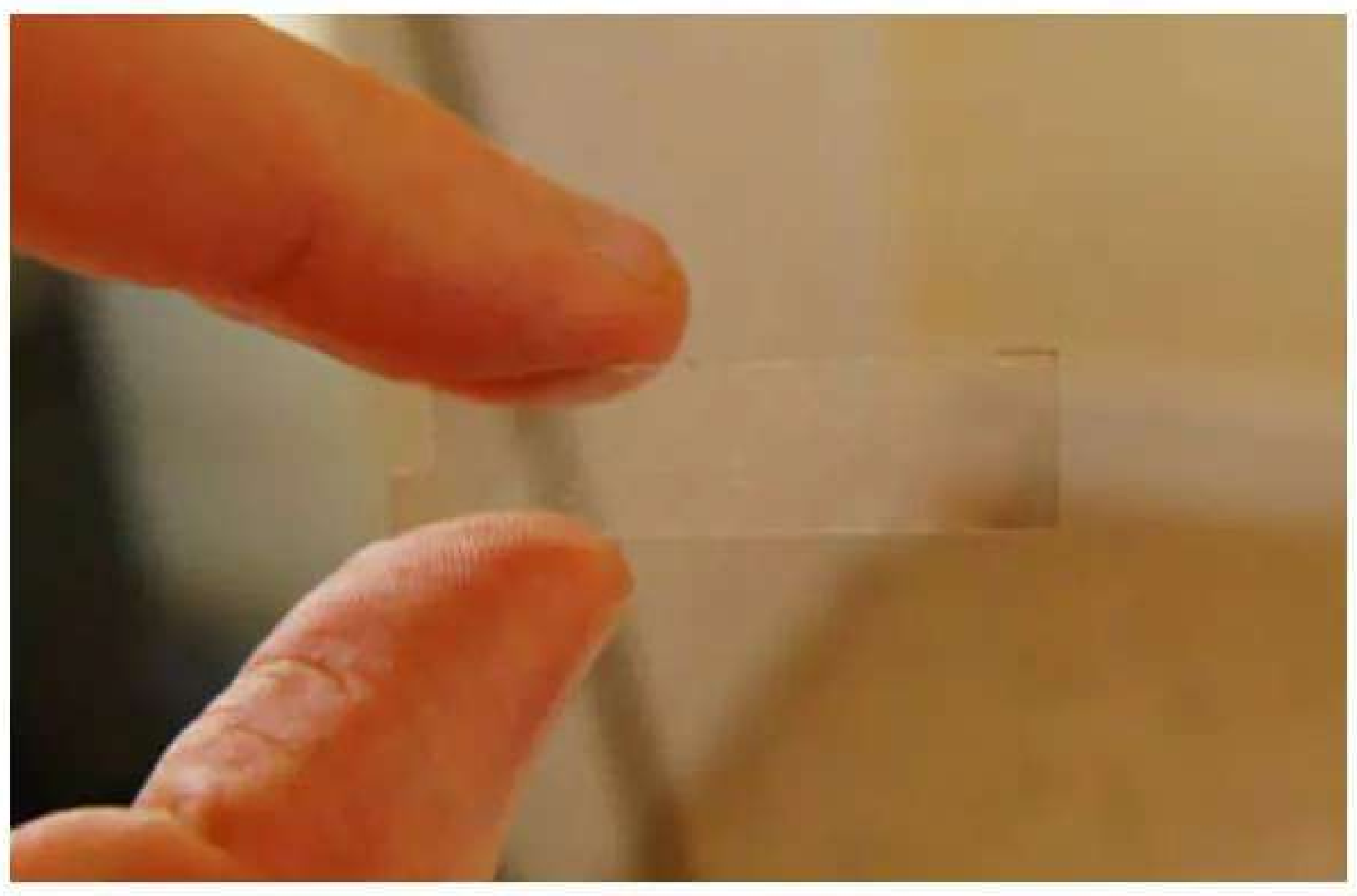


Fig.42: a sheet of capacitor-grade Indian ruby mica (Charcroft).

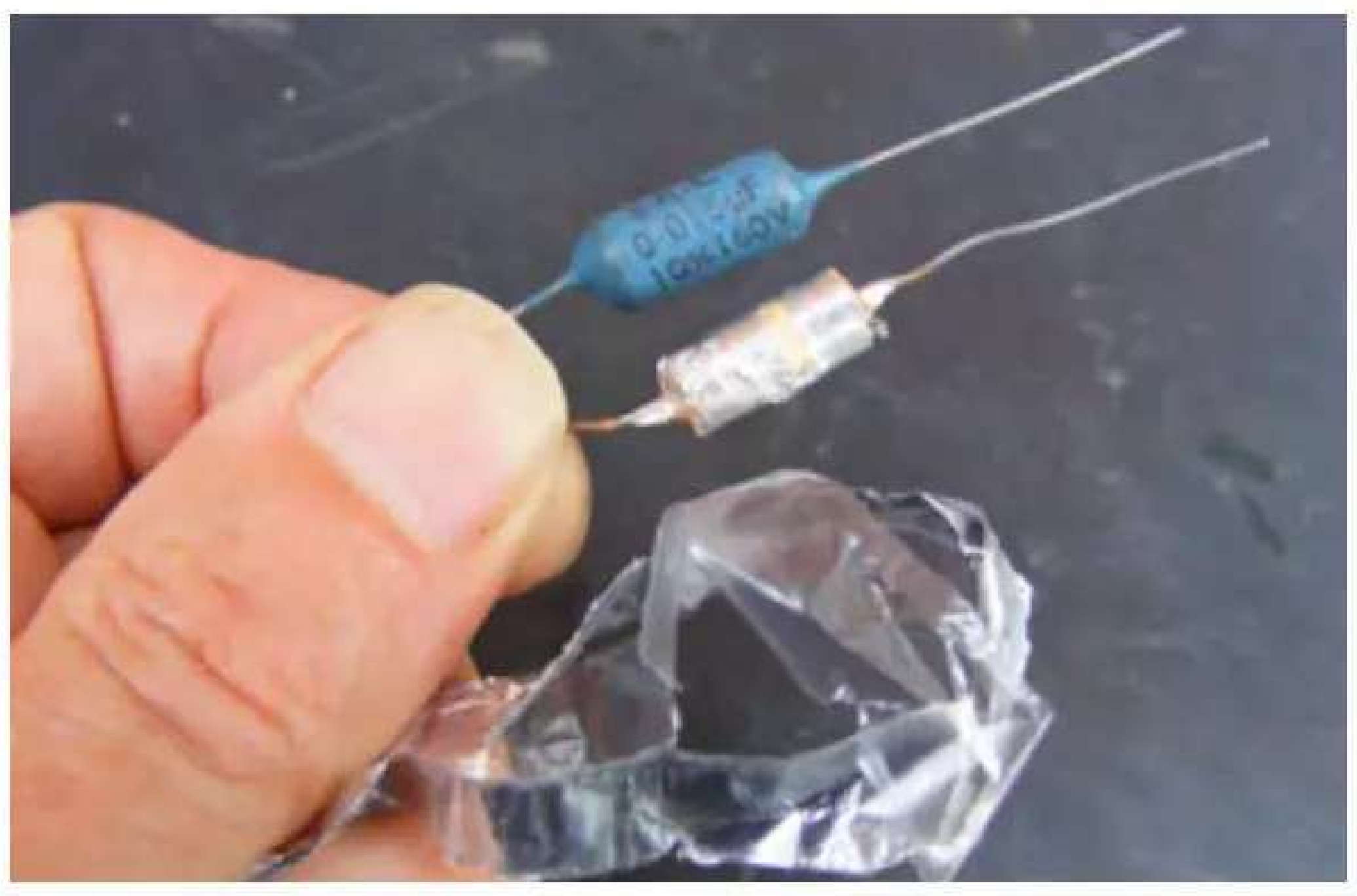


Fig.43: a polyester foil capacitor being unwound.

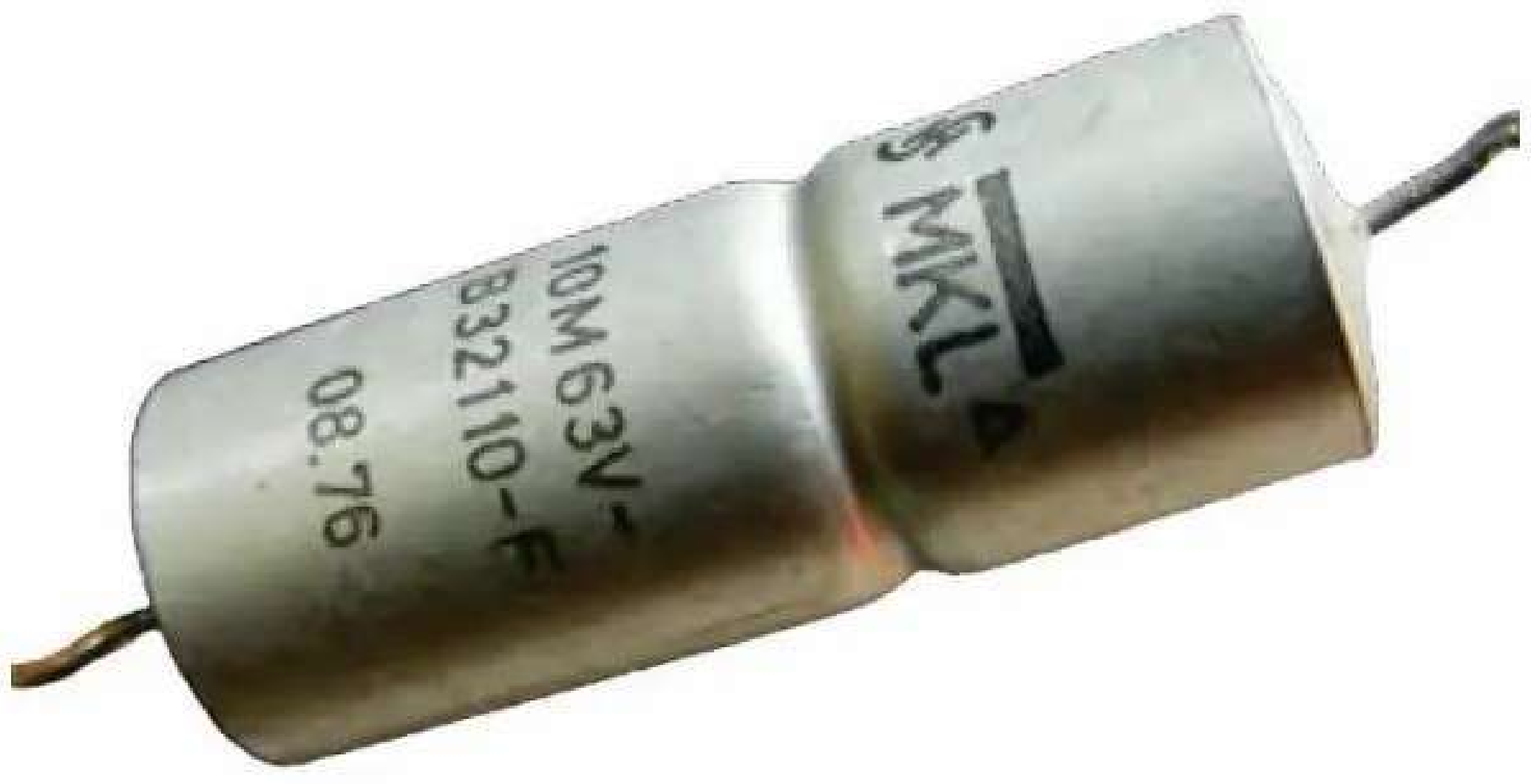


Fig.44: an unusual cellulose acetate (MKL) cap, the first plastic film type.

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cause hysteresis-induced distortion in some high-current situations. It may also increase distortion from the contact inside the capacitor. There is now a market for the older versions of capacitors that used copper leads. Generally, high-quality audio capacitors have pure copper leads; you can check this using a magnet (it will stick to steel but not pure copper).

Acoustic resonance

A subtle effect in film capacitors is where the capacitor has its own acoustic resonances that are driven by and contaminate the original signal. The effect is usually in the 10-20kHz region and can occur in large plastic-film capacitors used in passive crossover tweeter sections.

I had it once with some Wima MKB3 polycarbonate capacitors (Fig.36). You could hear an acoustic metallic "ting" around 7kHz when you tapped the metal case. Apparently, there were two versions of this capacitor; one was epoxy resin filled and one that had free air inside. The epoxy ones were non-resonant and sounded better.

Mordaunt Short used to parallel two capacitors in their tweeter crossovers so one would dampen the resonance of the other. It's always better in audio to have multiple low-level resonances than a high-amplitude single one. These effects are only audible to skilled listeners with good high-frequency hearing, though.

Encapsulation

Most capacitors are dipped in epoxy powder and then heated (see Figs.37 & 38). Higher-quality units are epoxy potted. Fig.39 shows a special potted 80nF mica capacitor made by Charcroft.

All capacitors are susceptible to moisture ingress, especially paper and solid tantalum types. It is possible to obtain glass/metal hermetically sealed tantalum types at high cost.

However, there are plenty on the military surplus market. Because they are hermetic, they are usually perfect, even if they have been stored since the Cold War.

Paper and mica capacitors were traditionally impregnated with silicone wax to improve stability and moisture resistance before encapsulation. This has to be done in a vacuum chamber to remove bubbles (see Fig.40).

Audio ranking

Here's a list of capacitors from best to worst for audio work.

- Teflon (PTFE), K=2.1. These types are very rare and expensive. Teflon is still used for trimmer capacitors.
- Polystyrene foil (FKS). K=2.56. Easily melted by soldering and difficult to source, but still the best for audio filters such as RIAA equalisers and active crossovers. Typically they have a 1-2.5% tolerance.
- NP0/C0G class 1 ceramic, K=30. These are good for HF compensation capacitors for amplifiers. They typically have 2–5% tolerance with a small size, excellent for SMT up to 10nF. Sometimes marked with a black paint band across the top.
- Polypropylene foil (FKP), K=2.2. They are good for values up to 47nF.
- Metallised polypropylene (MKP; see Fig.41), the standard quality plastic film capacitor type for audio. Note that there are fakes around. I've unwound a

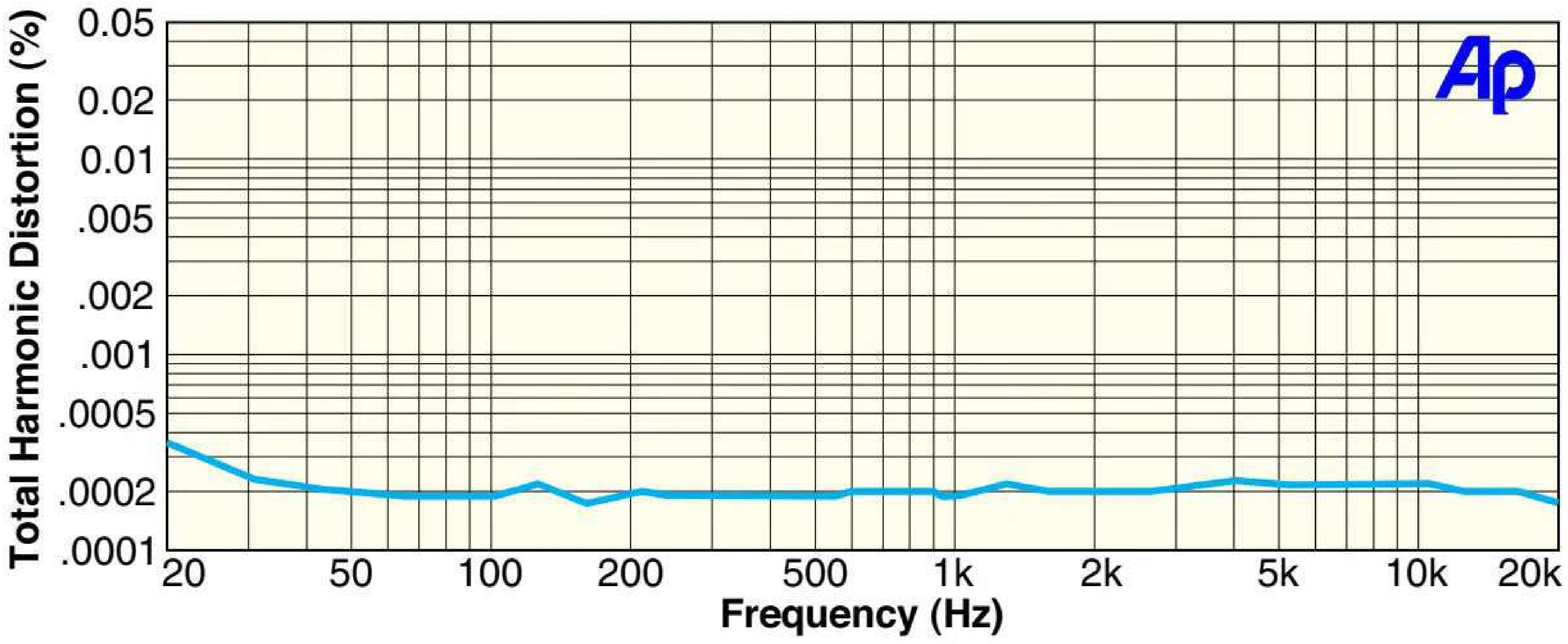

Fig.48: the very low distortion a 100µF 35V bipolar Suntan electrolytic @ 8V RMS.

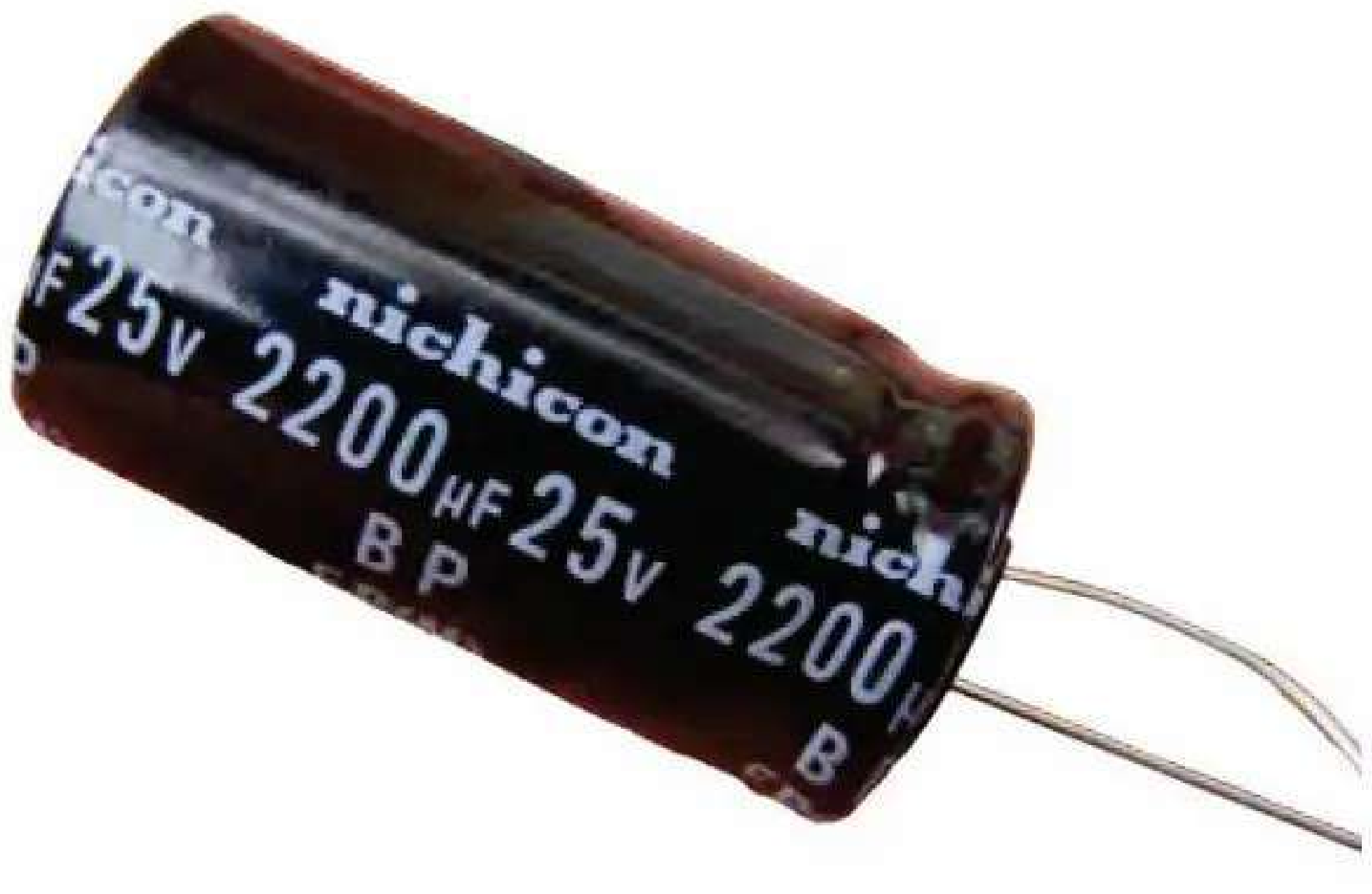

Fig.49: Nichicon make excellent bipolar electrolytics. This one is suitable for speaker coupling.

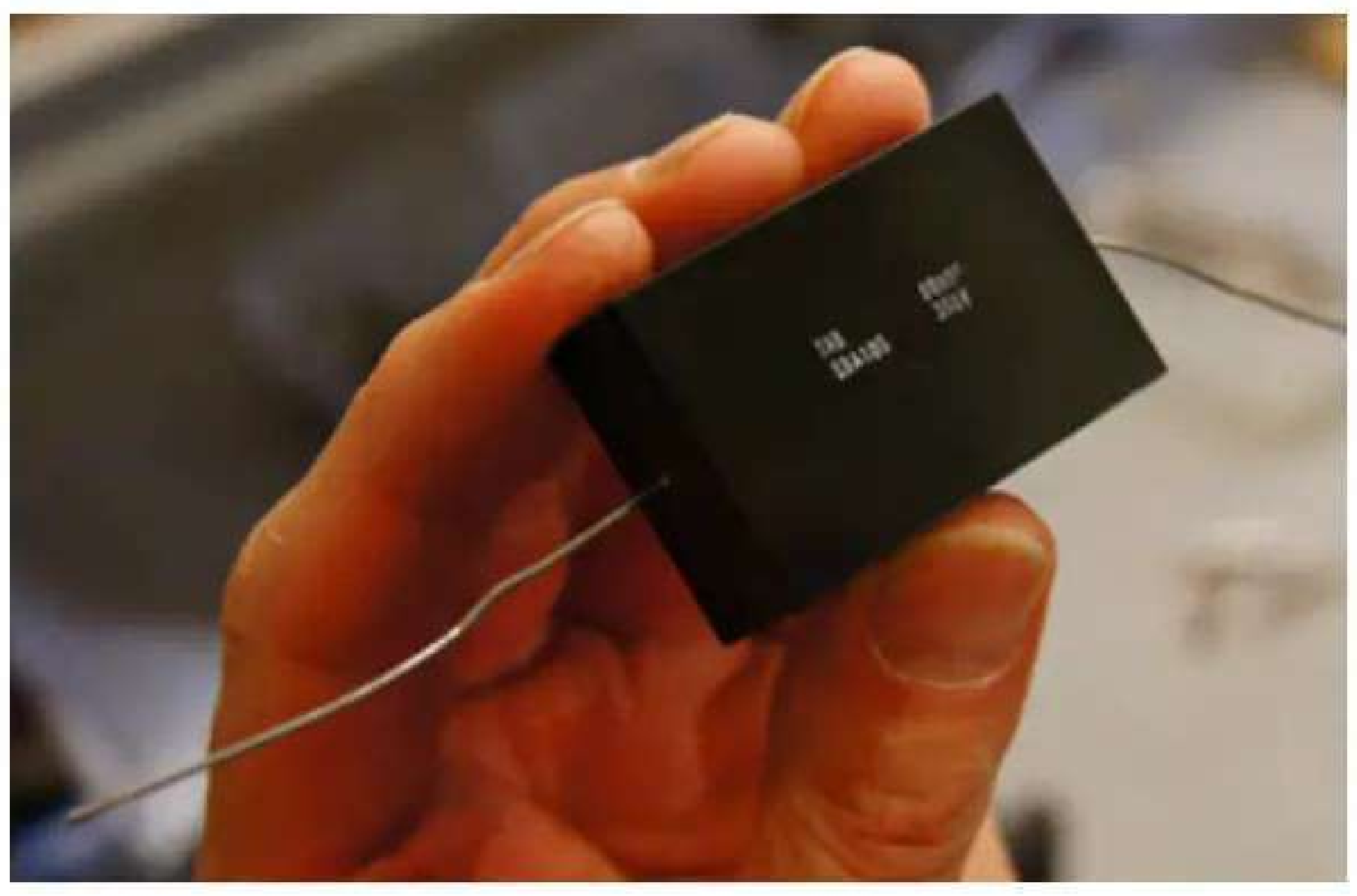

Fig.39: or this big ‘special’ 80nF silvered-mica capacitor was potted in epoxy resin.

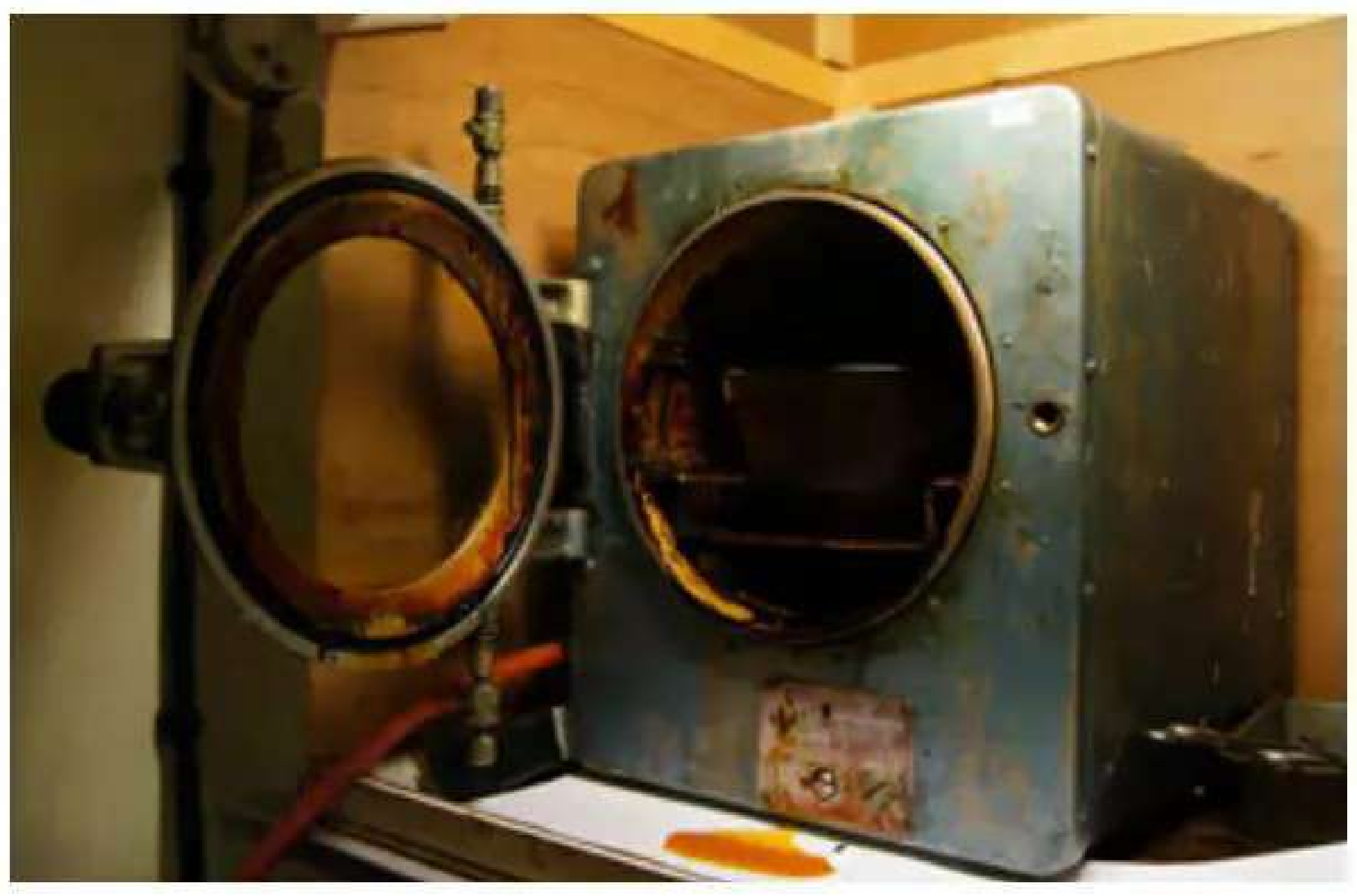

Fig.40: top-quality capacitors are wax impregnated in a heated vacuum chamber to exclude air bubbles.

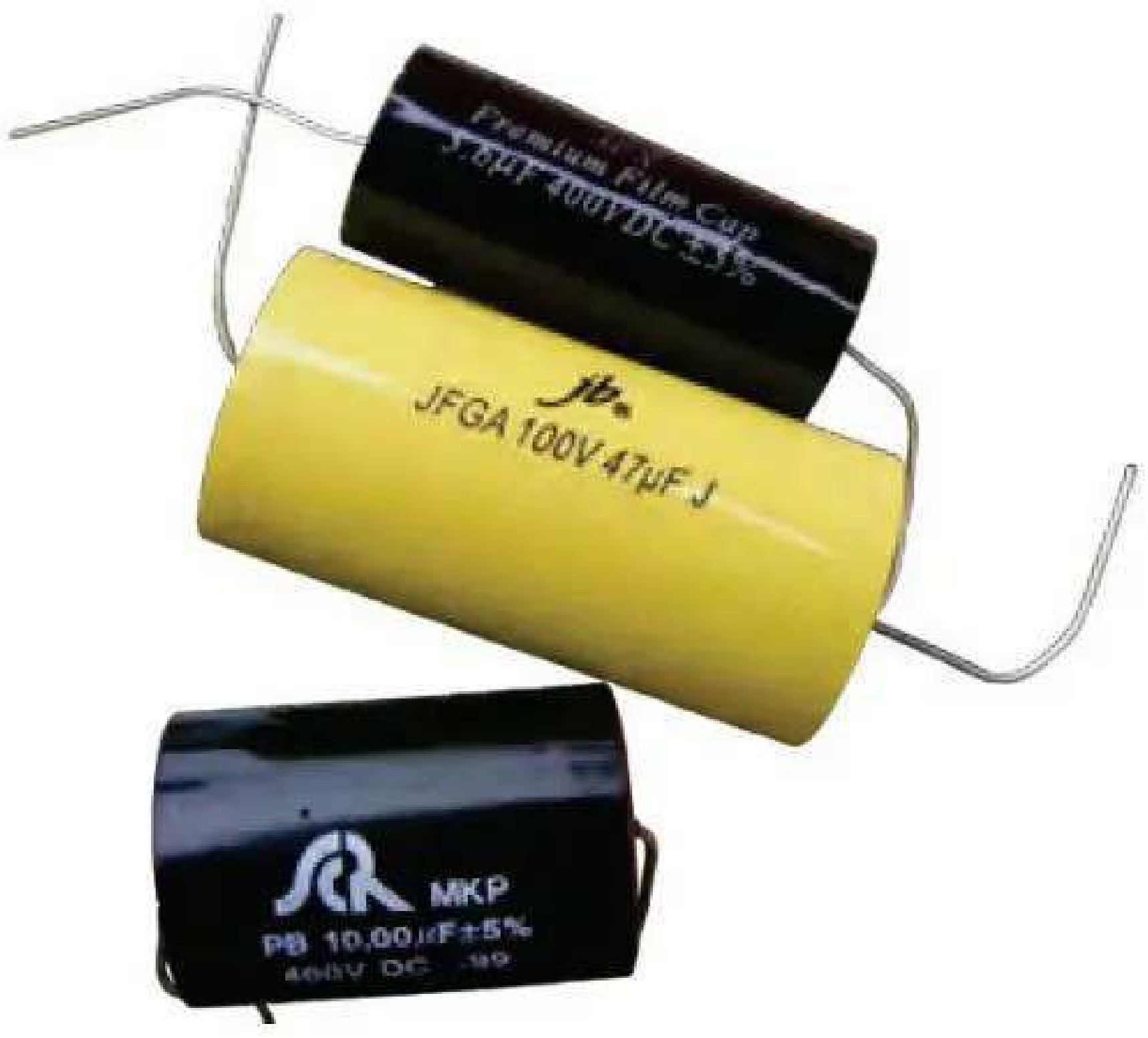

Fig.41: standard audio grade polypropylene capacitors, as used in speaker passive crossovers.

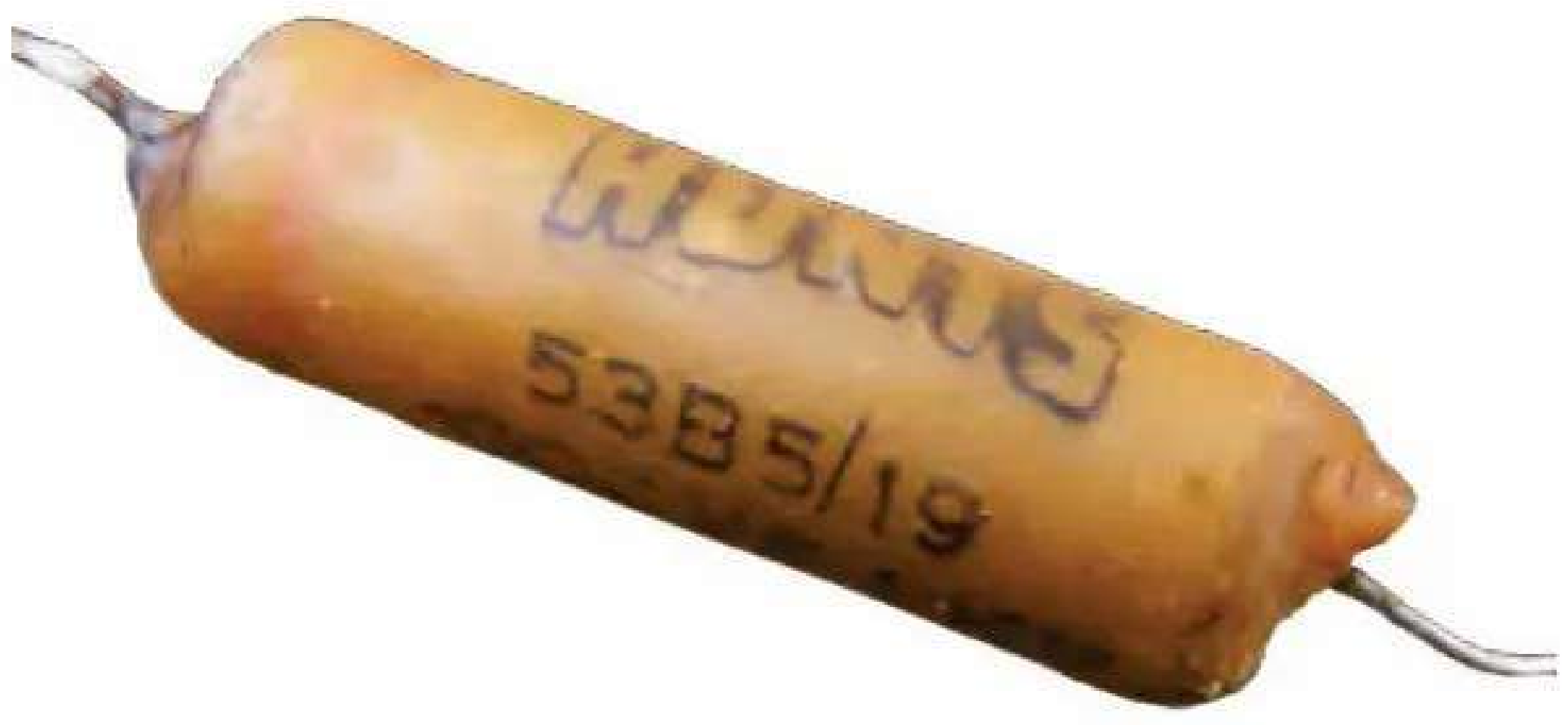

Fig.45: a wax-covered paper foil capacitor popular in 1950s valve gear.

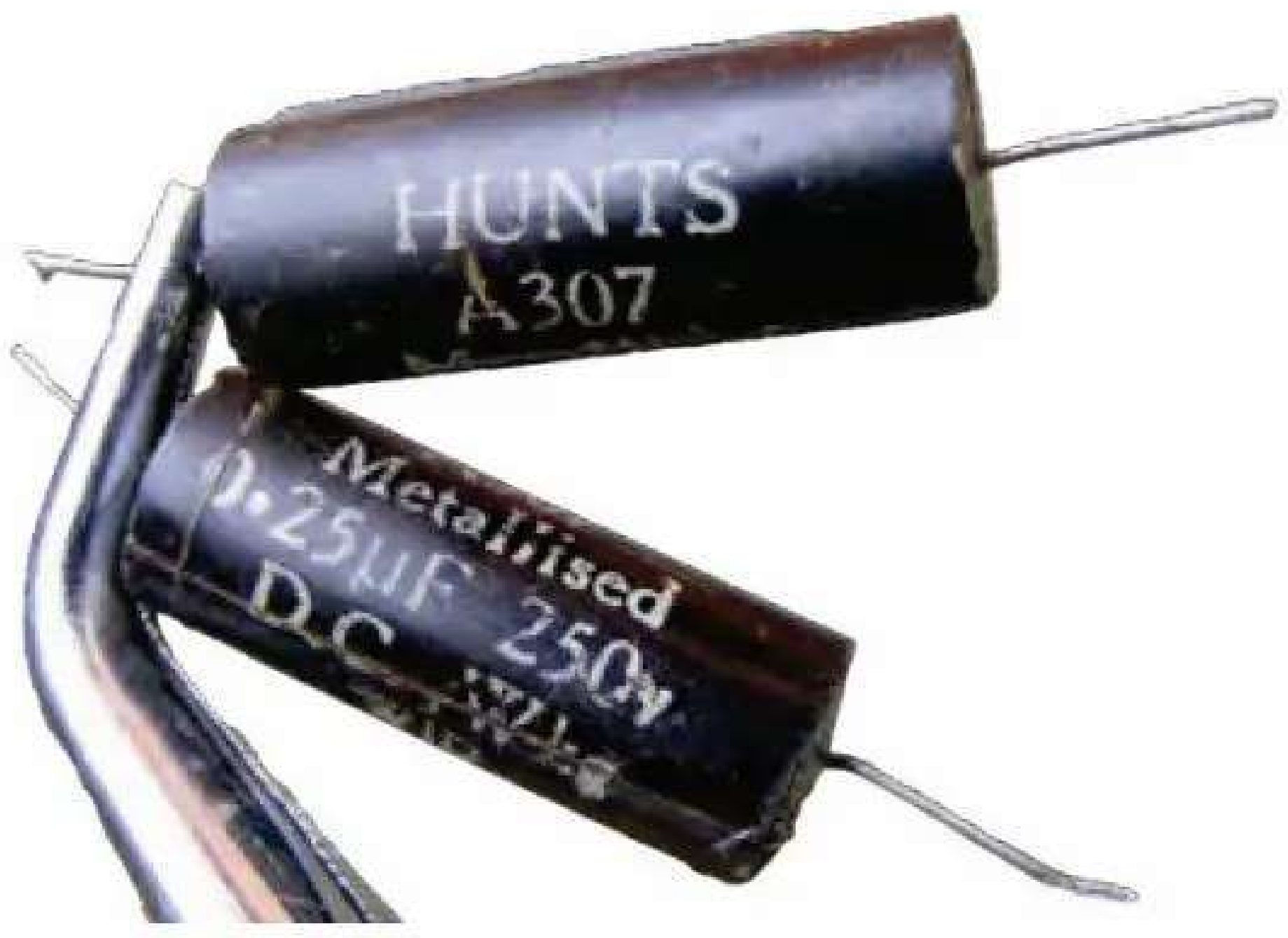

Fig.46: old metallised paper capacitors. These black Hunts types are unreliable.

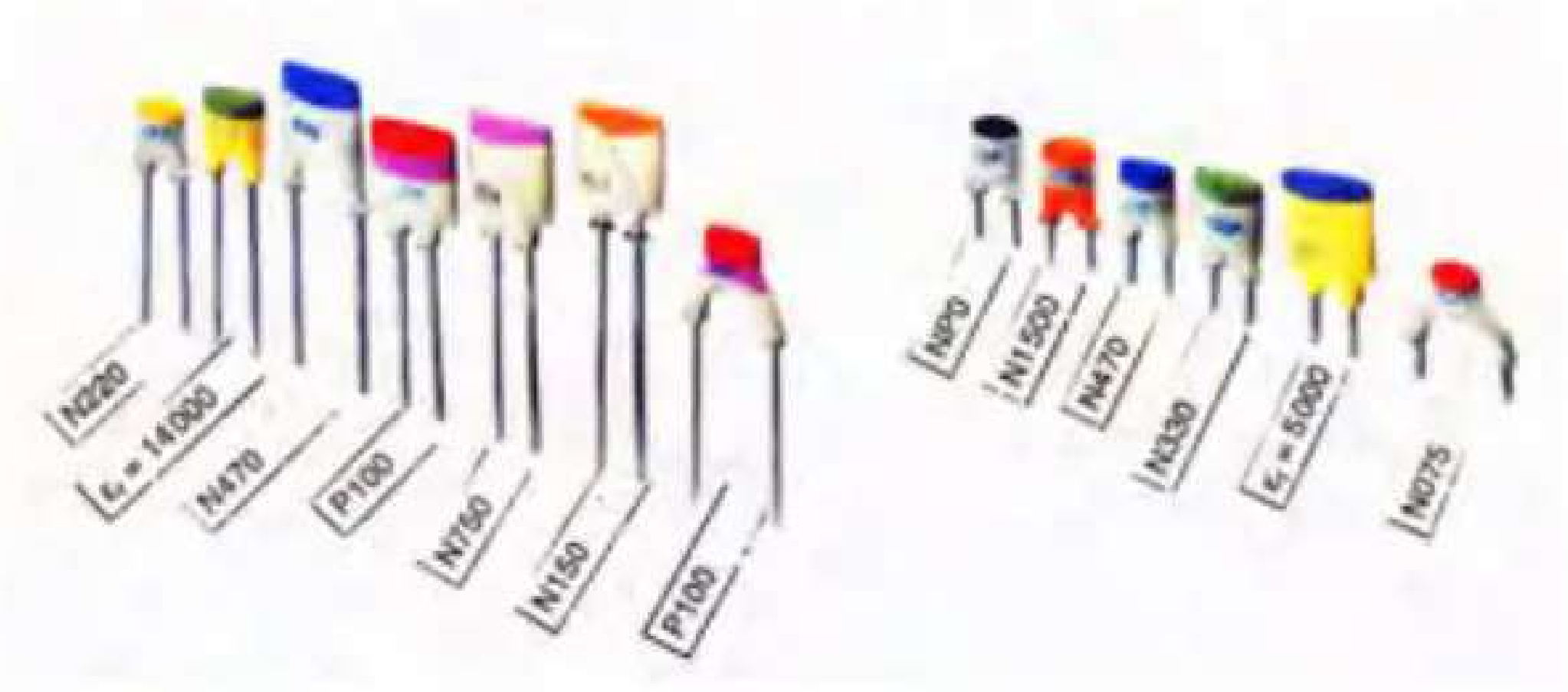

Fig.47: Philips ceramic plate capacitors are colour-coded by dielectric type.

few “audio polypropylene” capacitors and found a standard polyester type inside. You can tell as polypropylene is soft and polyester is hard.

- Mica (Fig.42), K=6. A proven natural material for capacitors over 100 years. Very stable, with 0.1% tolerance available, but expensive.
- Polycarbonate (MKC). Used in original BBC LS3/5A crossovers. A more stable version of polyester with a higher dielectric strength. K=3, and lower losses above 20kHz. They are no longer made.
- Polyester (FKT/Mylar/PET), K=3.2, used for soft drink bottles and the first practical plastic film dielectric. It was developed by German chemists during the Second World War to deal with the capacitor paper shortage. They are smaller than paper foil capacitors by 15%. Fig.43 shows a polyester foil capacitor being unwound.
- Metallised polyester (MKT). The general low-cost audio capacitor. 50% smaller than foil types.
- Glass, K=5.6. The original dielectric used for Leyden jars. They are radiation-resistant and have a very wide operating temperature range.
- Lacquer film/cellulose acetate (MKL), K=4. Like paper, but very small and radiation resistant. Were only made by Siemens (see Fig.45).
- Paper in oil. Used by valve enthusiasts. The viscously damped soft paper gives a very low mechanical resonance.
- Impregnated paper-foil K=4. Often only dipped in wax, as shown in Fig.45. They absorb moisture, so they can be leaky.
- Metallised paper. Avoid the Hunts (Fig.46) and Rifa types like the plague.
- Mixed dielectric, made by Dubilier. They look like a blue-and-white chewy sweet. They have the low leakage of polyester with the sound quality of paper. Polyester is the main dielectric between the foils. They were popular in old TVs.
- Medium-K ceramic. Unstable, with a large negative temperature coefficient. Philips used to make a whole range of square ceramic plate types with differently coloured tops (see Fig.47). Orange was -150ppm/°C and violet was -750ppm/°C. I used them for temperature compensation in tuned circuits with ferrite inductors.

Esoteric electrolytics

- Wet aluminium bipolar types are the cheapest option for a low-distortion electrolytic (Fig.48). Nichicon's UEP series (Fig.49) capacitors are excellent.
- Cerafine by Elna, with a paper separator impregnated with fine ceramic particles. The distortion specification is -120dB at 10kHz & 0.1A. Distortion does increase at low frequencies in the normal way (see Figs.50 & 51).
- Slit-foil electrolytics (Fig.72). These capacitors are made by Kemet and designed by Denis Morecroft of DNM. They are the best speaker coupling and main reservoir caps. The foils are slit to reduce circulating eddy currents, lowering distortion (in theory). They were used in Mission Cyrus II amplifiers. There is even a four-terminal version, the T-network capacitor, with internal connections at optimum points on the foil for minimum rectifier spike induction, giving 10dB less HF noise.
- Silmic branded audio electrolytics with special silk paper made by Elna. They have lower distortion than Cerafine types.

Cerafine
100µF 100V

Fig.50: a pair of Cerafine low-distortion electrolytic capacitors.

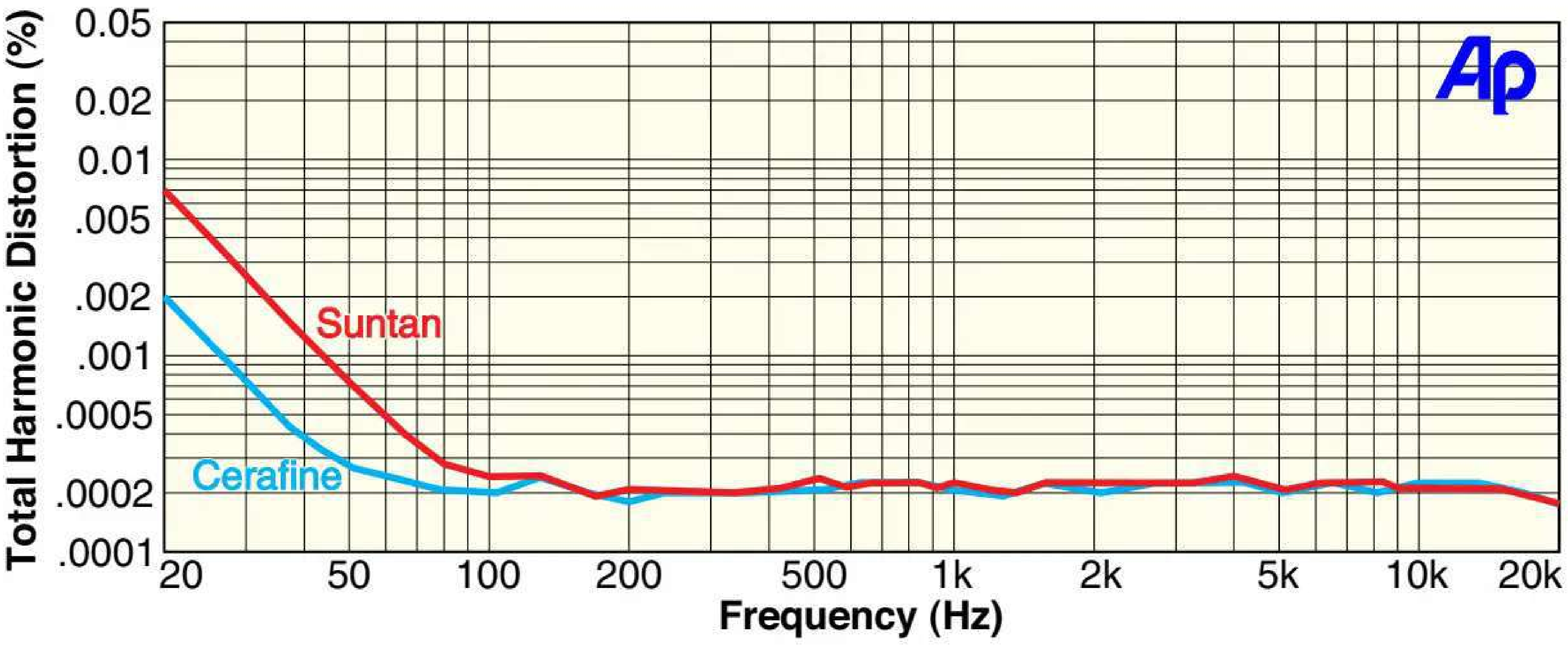

Fig.51: the distortion of Cerafine versus normal (Suntan) electrolytic caps.

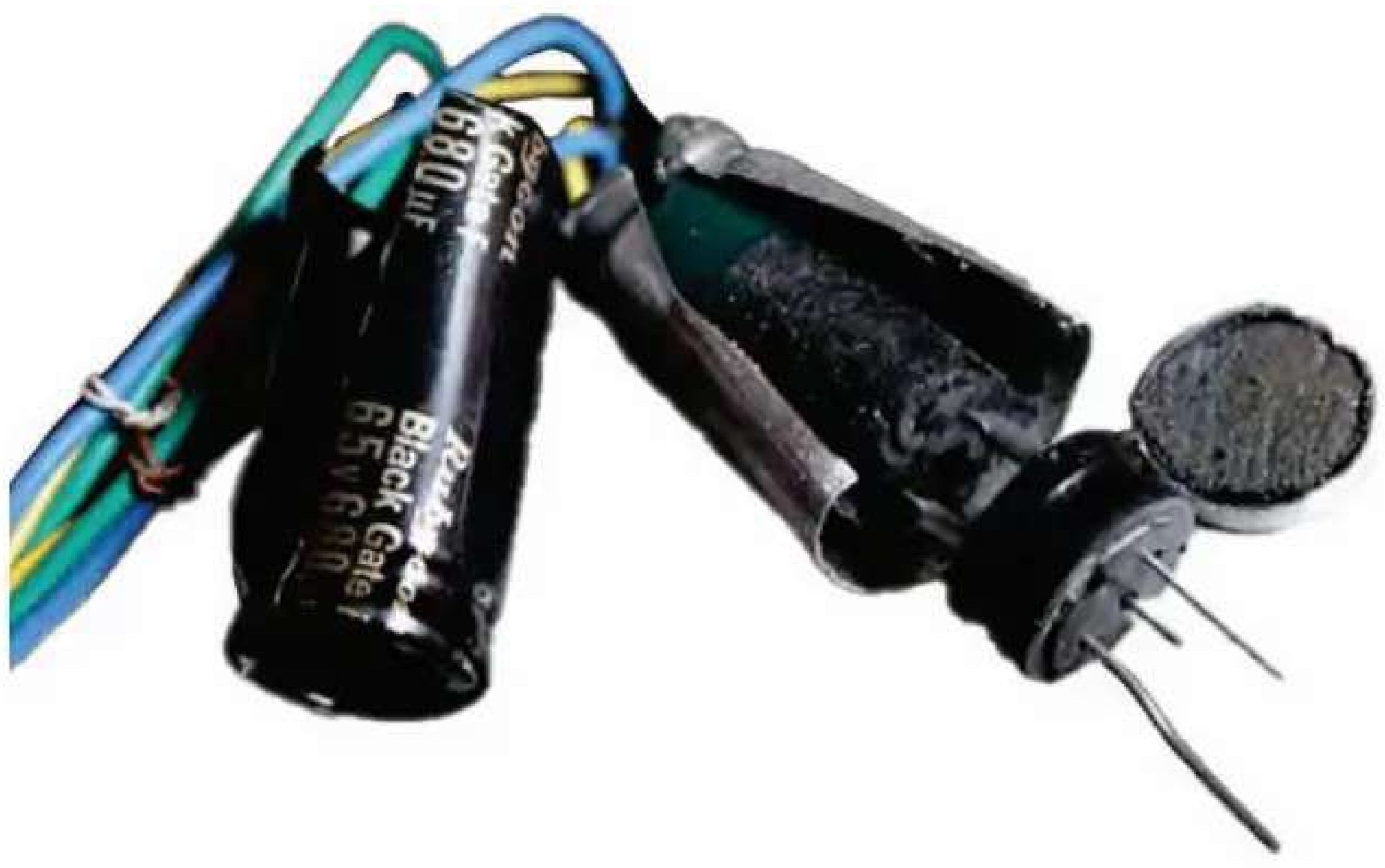

Fig.52: a Black Gate electrolytic cut open (from Jimmy's Junkyard). Note the black carbonised winding.

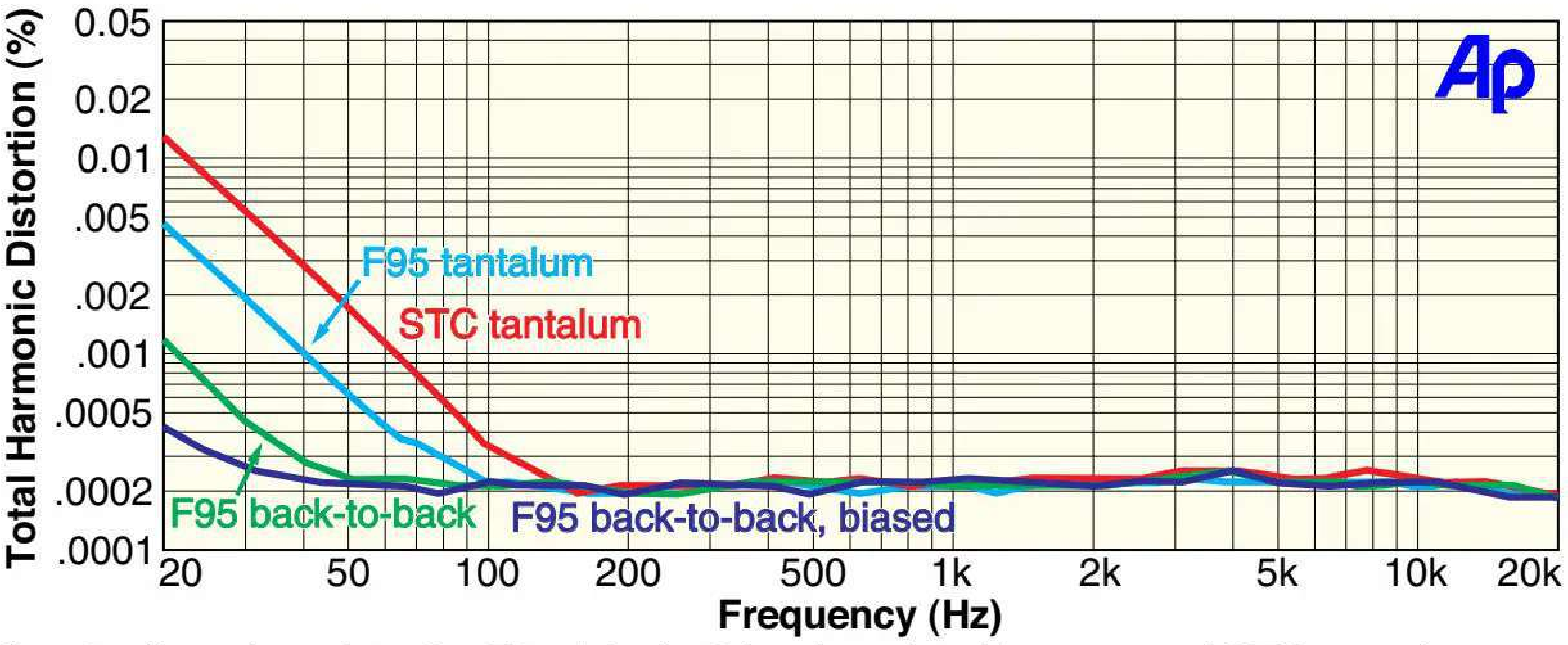

Fig.53: distortion plots for F95 (single & back-to-back) vs normal STC tantalums.

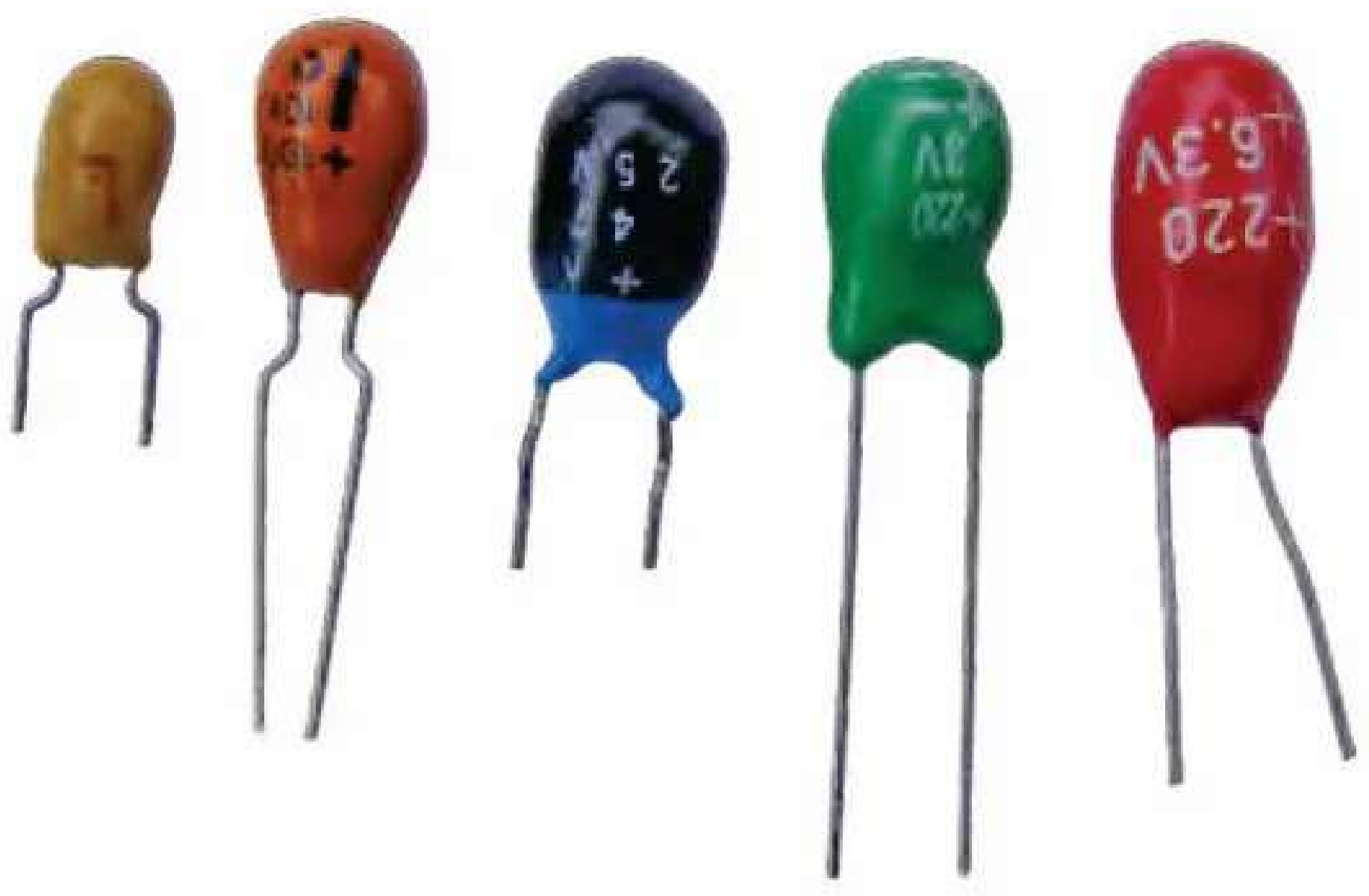

Fig.57: solid tantalum bead types. They can suffer from moisture ingress. The type on the right seems prone to failure.

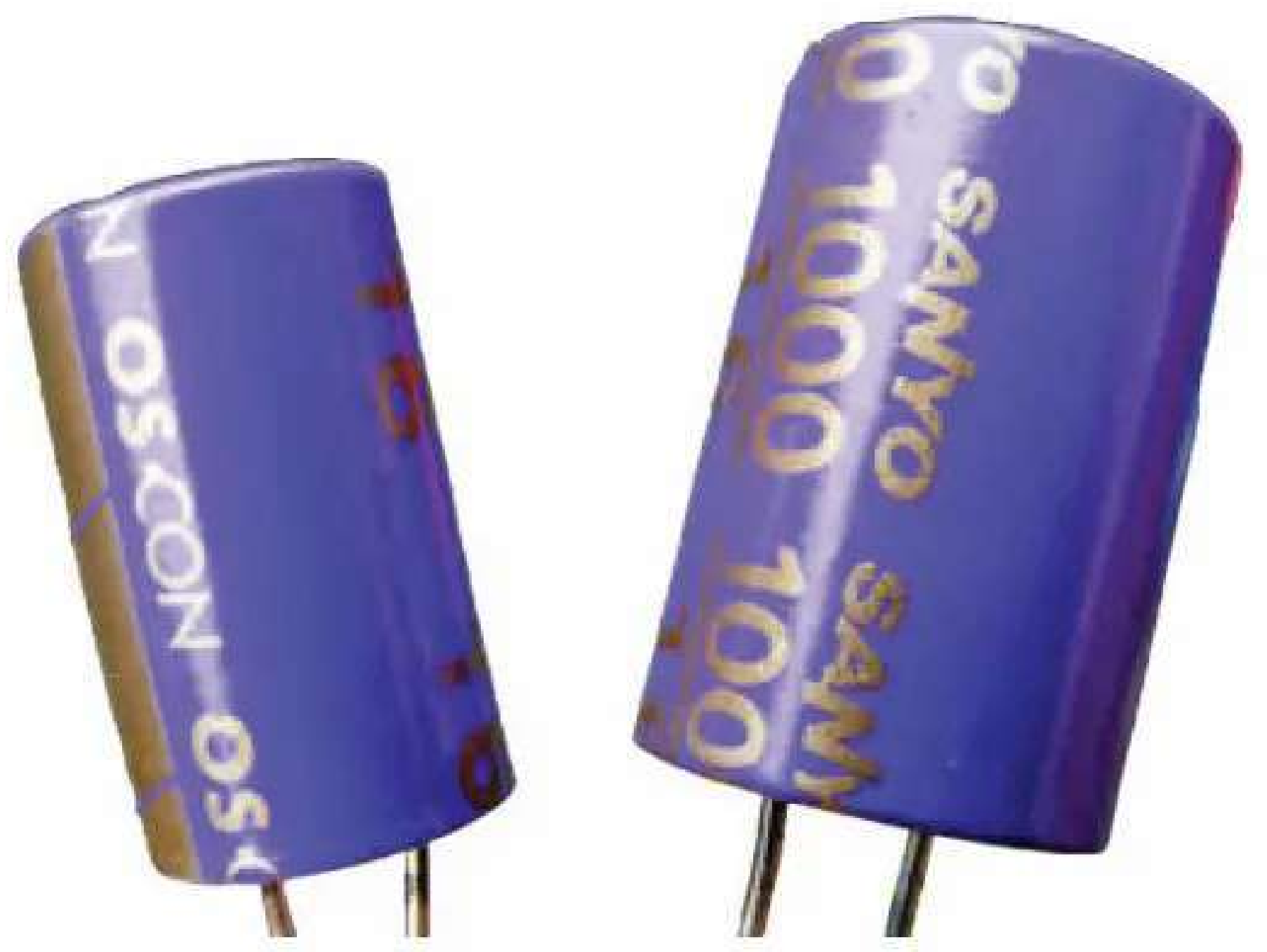

Fig.58: Oscon solid-aluminium polymer capacitors from the SG audio series. These have an epoxy-sealed base.

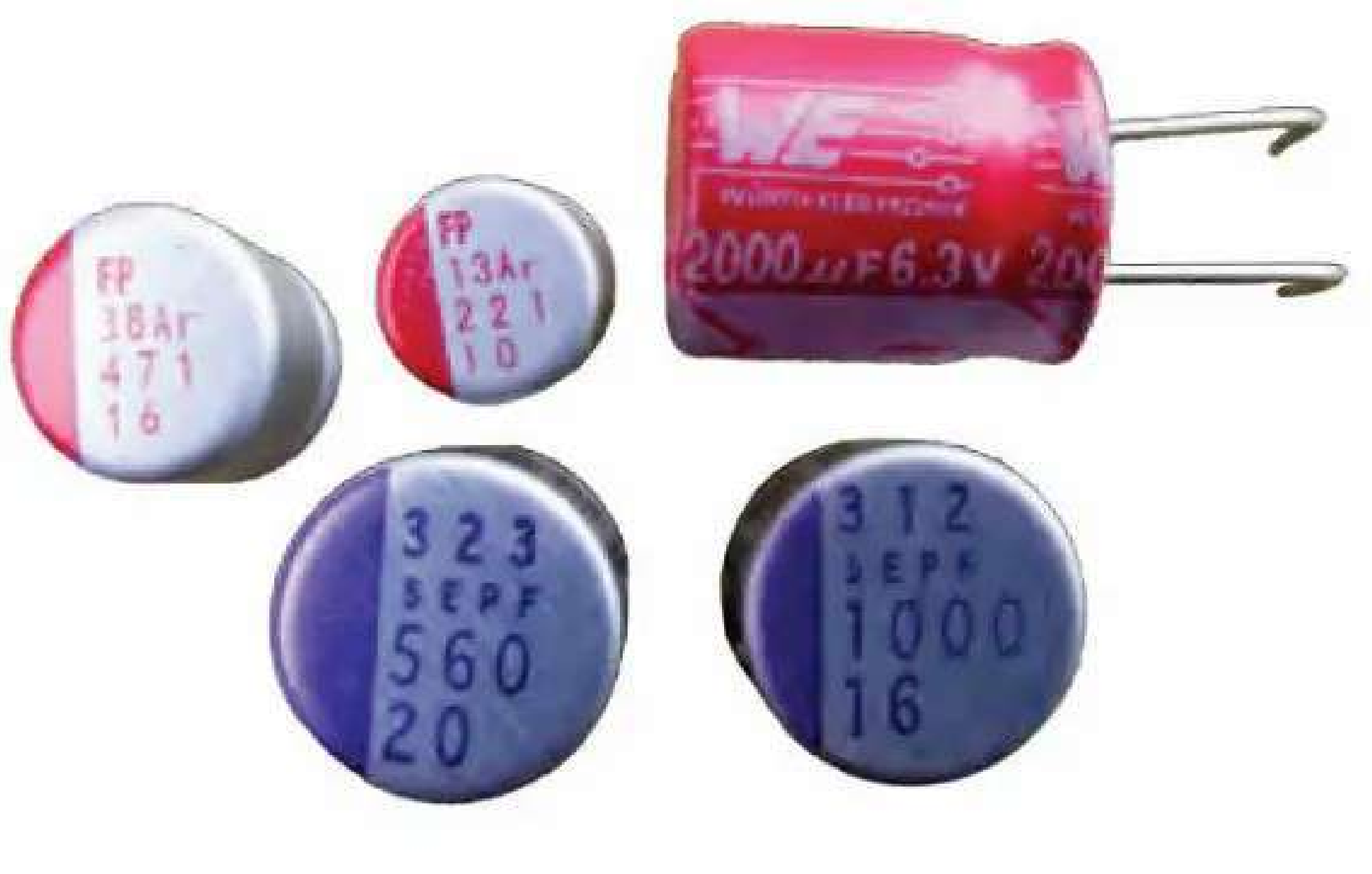

Fig.59: Panasonic SEPF, Nichicon FP and Würth Elektronik solid polymer electros. These all have rubber bungs.

- Black Gate by Rubycon, an audio electrolytic with graphite-impregnated paper separator. The paper winding looks black if you open them up (Fig.52). They had the best distortion, down to -150dB at 10kHz as per the data sheet. They were discontinued in 2006. Audio Note Kaiseis are the new replacement, but their cost is beyond engineering comprehension.
- Audio F95 tantalum by AVX (Fig.54). Sadly they are only available in SMD packages, with a quarter the distortion of conventional solid tantalums, as shown in Fig.53. Apparently, the lower distortion is possible due to a thicker dielectric and frameless design. When biased, distortion drops further to 0.001%. These are the electrolytic caps to use for top-quality, high-reliability audio. At £1.92 a pop, they are expensive, but very small.
- Metal-cased solid tantalum (Fig.55). These seem to last forever. They are made mainly by Kemet today, eg, the T110, CSR13, Vishay CTS13 and STC TAA series. They are very expensive at over £5 new, but cheap on the surplus market. These were used in Nagra tape recorders and avionics.
- Wet tantalums are very rare, and the most expensive, an eye-watering £40 plus. Good for high voltages. Distortion should have been very low, but when I measured the few military surplus samples I had, such as the bipolar 15μF 100V Stantalum in Fig.56, it was worse than a normal solid tantalum. I ran it at 8V RMS into a 600Ω load, whch gave 0.15% distortion at 20Hz.
- Tantalum beads (Fig.57). Kemet's and Union Carbide's are excellent. The old ITT/STC red and blue types have poor reliability; their later yellow ones were better.
- Polymer tantalums have lower ESR and don't ignite if abused, like manganese oxide tantalums. Their distortion may be lower, but I haven't checked yet.
- Oscons, the first solid polymer electrolytics, developed by Sanyo but now made by Panasonic. Distortion is still high, but the ESR and hence decoupling ability was in a new league. They did an audio version shown in Fig.58, the SG, which had copper leads, but the distortion was still slightly higher than solid tantalum. However, their CV product went up to a higher level than tantalum types, allowing more wet electrolytics to be eliminated from high-reliability designs.
- Solid aluminium manganese dioxide electrolytic, a Philips invention to get around the high cost of tantalum. They have too much leakage for audio coupling use, and high ESR, although this is beneficial for regulator output damping. They were the most reliable electrolytics, with high reverse voltage ratings, but are now obsolete.
- Generic solid polymer electrolytic (Fig.59). Their very low ESR makes them fantastic for decoupling, but can cause regulators to oscillate. To stop this, I've had to insert a resistor in series, but this does allow the ESR to be optimised and defined. Their leakage is four times higher than wet aluminium, so it says "not suitable for coupling" on the datasheet. Also, the distortion is twice as high as tantalums; back-to-back connection and biasing does not seem to help. They easily go short circuit if reverse biased. I've found them to be excellent emitter bypass and bootstrap capacitors in audio amplifiers, though.
- Hybrid polymers contain liquid electrolyte but it's impregnated with polymer particles, giving the low ESR of polymer but with the self-healing characteristic of normal electrolytics. They give the best of both types, but can still dry up. I bought a couple of samples of a Kemet hybrid polymer (Fig.60). These are built into the almost hermetically sealed Rifa-style axial case design with a metal end disc. They show great promise as a speaker coupling capacitor, being similar to Black Gate types but at an industrial price, rather than an "audiofool" price.
- Niobium oxide; although advertised as "Oxicap audio caps", these have higher distortion than tantalums. I think it is because the oxide's linear bandgap region is smaller.
- X7R multilayer strontium titanate ceramic, K=250. Use for decoupling only. Their distortion is ten times that of tantalum types.
- Z5U/Y5V barium titanate ceramic, K=600-1200. Use for decoupling only. Even worse distortion than X7R!
- Barrier-layer ceramic class 3, K up to 50,000. Introduced by Erie (Fig.61) in the 1960s for transistor decoupling. They often have very low voltage ratings, eg, 12V. They use a semiconductor effect to give high capacitance with a single layer but can't be made with multiple layers. The distortion is so bad that they are useful for audible demonstrations of capacitor distortion!

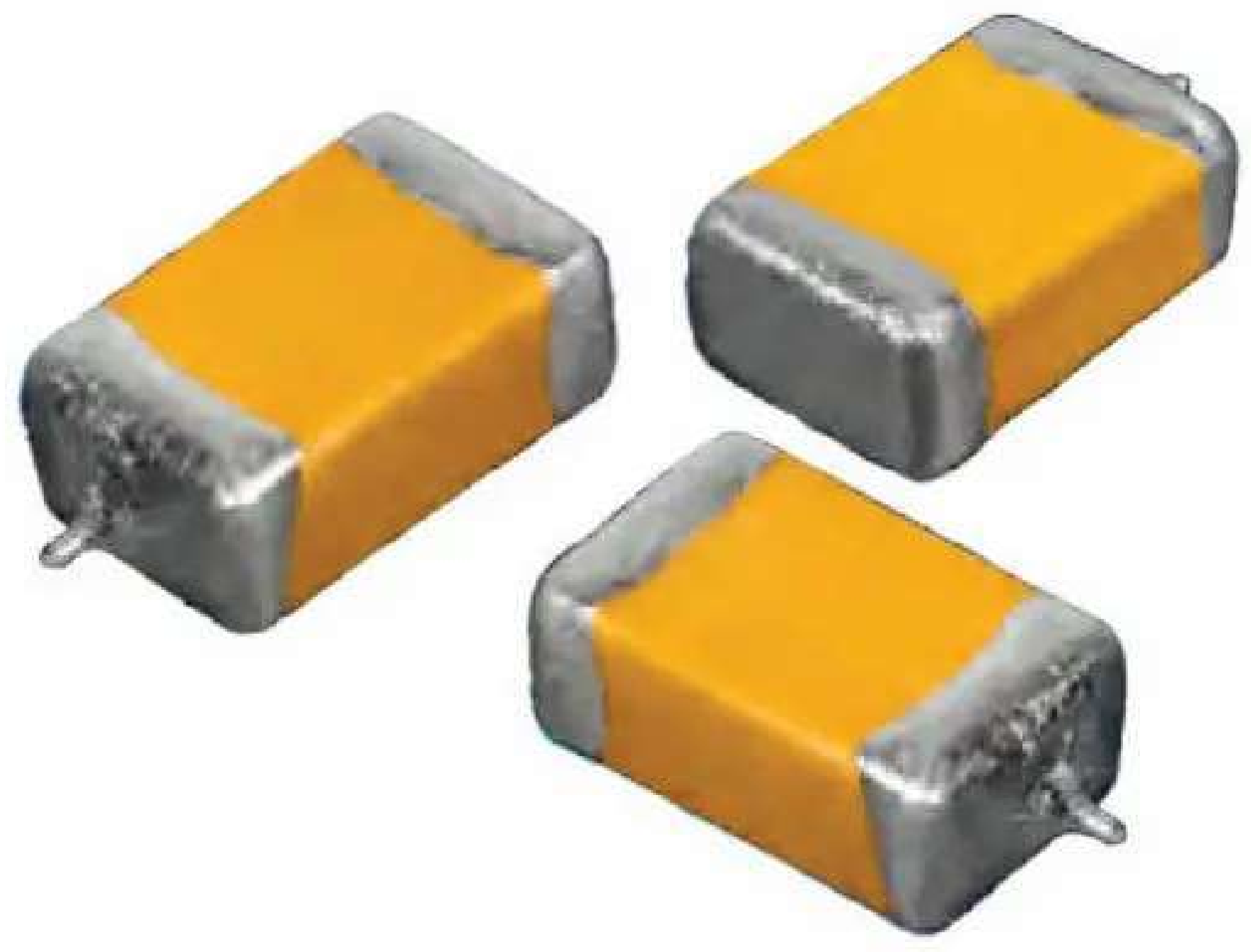

Fig.54: the F95 audio tantalum from AVX. The spike indicates the positive terminal.

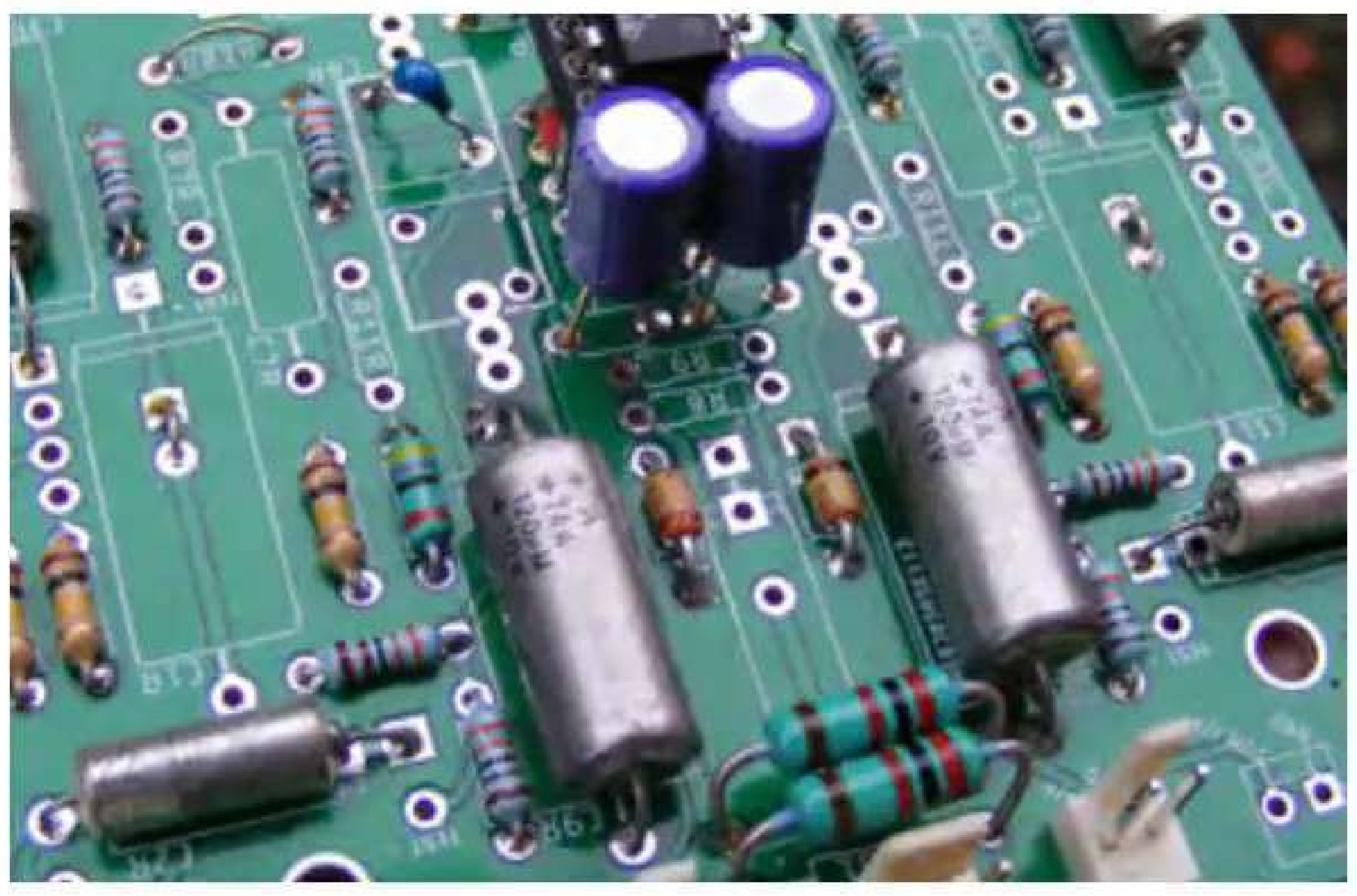

Fig.55: metal-cased hermetic tantalums from the ITT TAA series and Sprague 150D series; these last for decades.

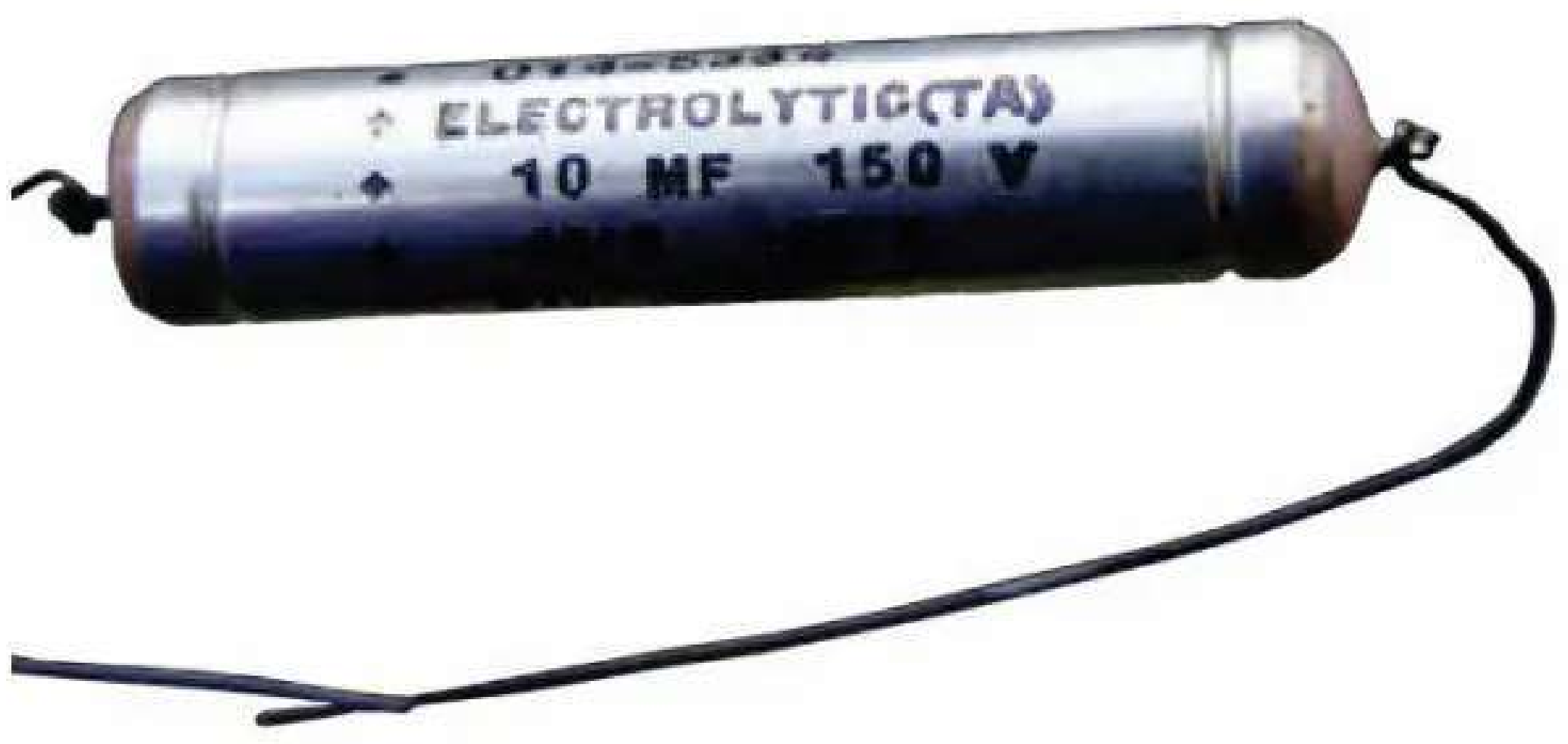

Fig.56: a 15μF 100V Stantalum bipolar wet tantalum by STC. I'm not sure what is so special about this type.

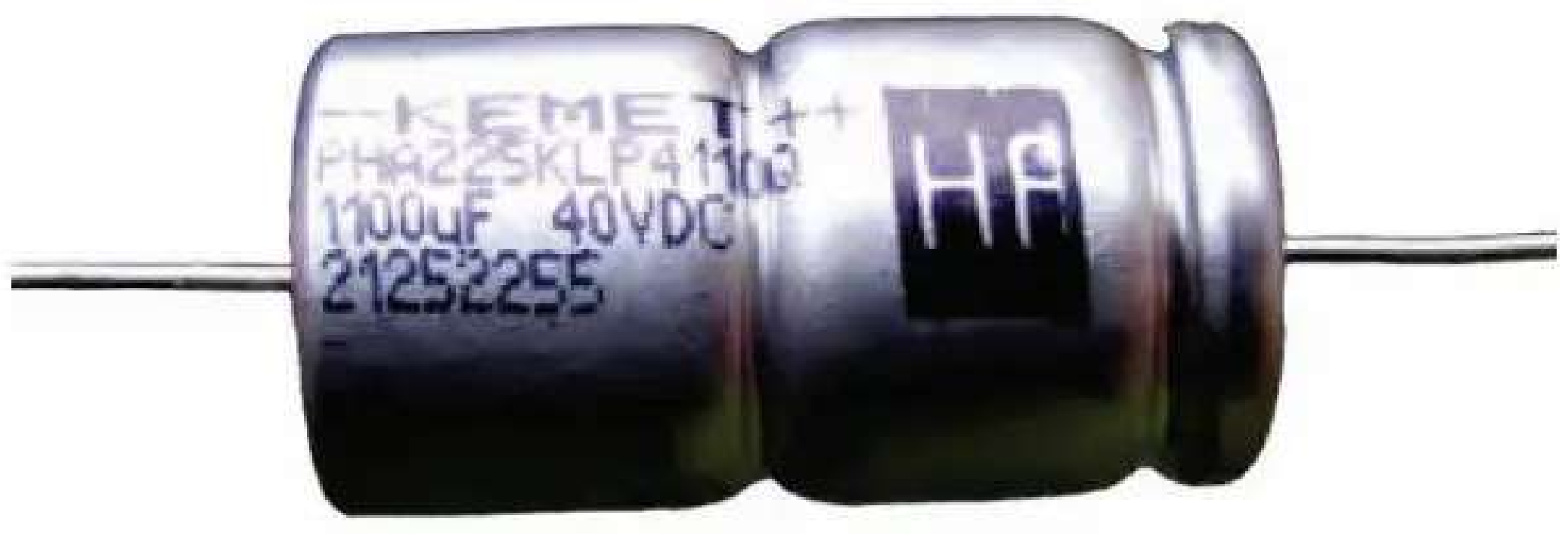

Fig.60: a Kemet hybrid polymer electrolytic, which has a very long life for wet technology.

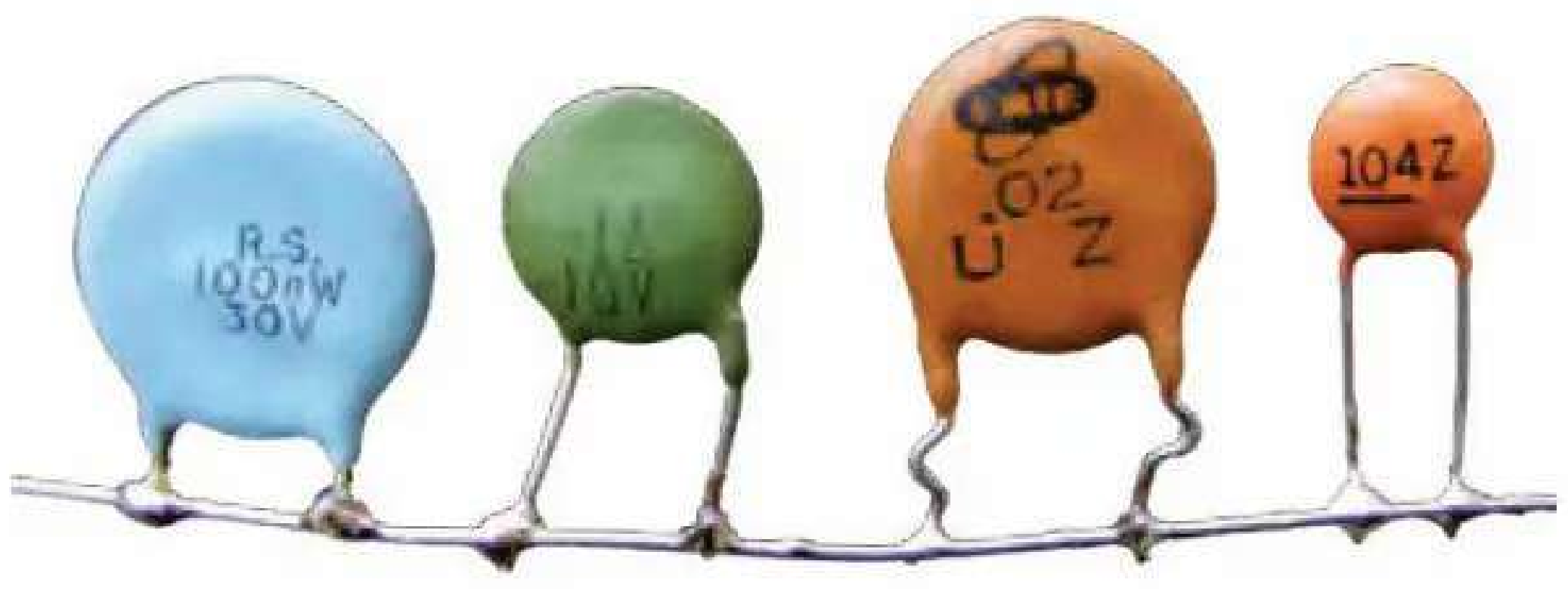

Fig.61: surface-barrier ultra-high-K ceramics with distortion in the order of 2%. Maybe useful for fuzz boxes!

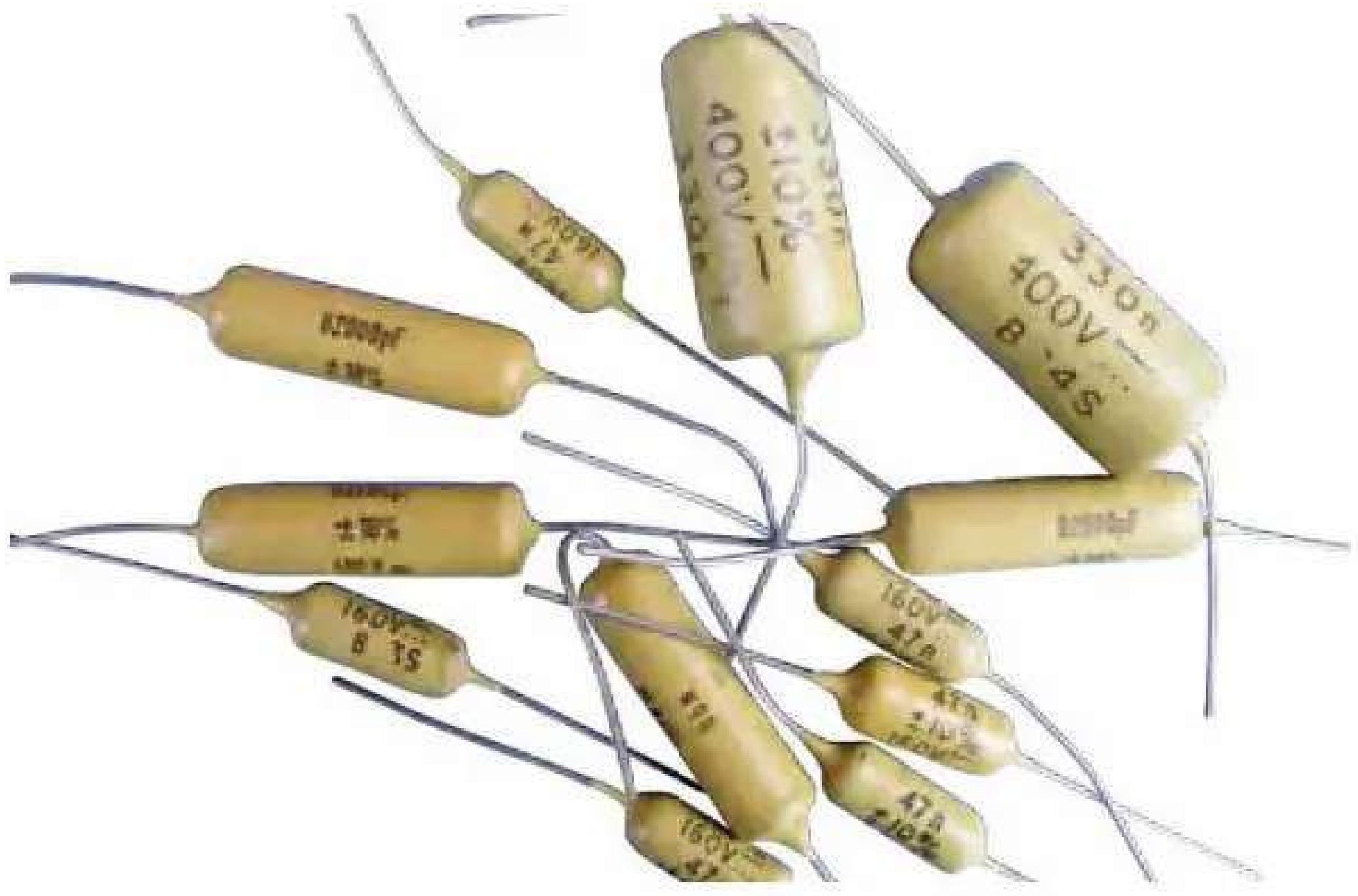

Fig.62: Mullard Mustard capacitors, officially called type C296. A valve amp maker's dream for the last 65 years!

Classic audio capacitors

Wima

The plastic film capacitor was a German invention pioneered by the Wima company, which is still going strong today.

Cambridge Audio uses them and they advertise that fact. The MKS2B051001N00JS 10μF 50V type is only 11mm wide and 16mm high with a 5mm lead pitch. However, it costs £3.50.

Mullard/Philips

This company had a huge following in capacitors. They later reduced production, handing it over to BC Components and Vishay. Their transparent shrink-wrapped blue electrolytics have always looked good, but go leaky if not polarised.

"Mustard caps"

These are Mullard's most famous capacitors (Fig.62), more like Dijon mustard rather than English. If you see these in an old amplifier or radio, do not replace them.

These axial polyester-foil capacitors never go wrong, apart from the leads falling off. They were used in the Philips Electronic Engineer kits and they were the first UK plastic film types, introduced in 1958. They are no longer made, highly collectable and I've seen a fight at a radio rally over them.

Tropical Fish, Mullard/Philips C280 capacitors

These are loved by guitar pedal builders and are so pretty they have been used as ear-rings. They are quite poor capacitors with fragile lead attachment, 10-20% tolerance and high distortion. I had an old Solina organ full of these and I got a thrill just looking at the PCBs.

They were introduced in 1962 and discontinued in 1984. They became the Vishay MKT368, which have a nice orange epoxy coating, a European version of the US "Orange Drop" made by Sprague.

"Black Beauties" and "Bumble Bees"

These are more nostalgia fest capacitors from US company Sprague, loved by guitar people. Retro Vibe have reissued the Bumble Bee for £33 (see Fig.63)!

The Fender Stratocaster forum showed a picture of a fake Bumble Bee cap (Fig.64), they are meant to be paper foil types. With a $5000 guitar, you want to maintain authenticity. Maybe I should start dipping my Chinese polyesters in paint to make "tropical fish" caps. They would be better than the real thing!

Mullard/Philips polystyrene

One of the most stable capacitors ever made, usually with a 1% tolerance. They are now banned by environmental laws because they used lead foil. This was, of course, their great advantage because the soft lead supporting the hard brittle polystyrene avoided the cracking that occurs when aluminium foil is used.

The lead also ensured an excellent soldered connection on the end of the winding in comparison to the difficulties in soldering the wire contact to an aluminium electrode.

The original yellow "Mini Poco" version (Fig.65) used copper leads and is highly sought after, along with its

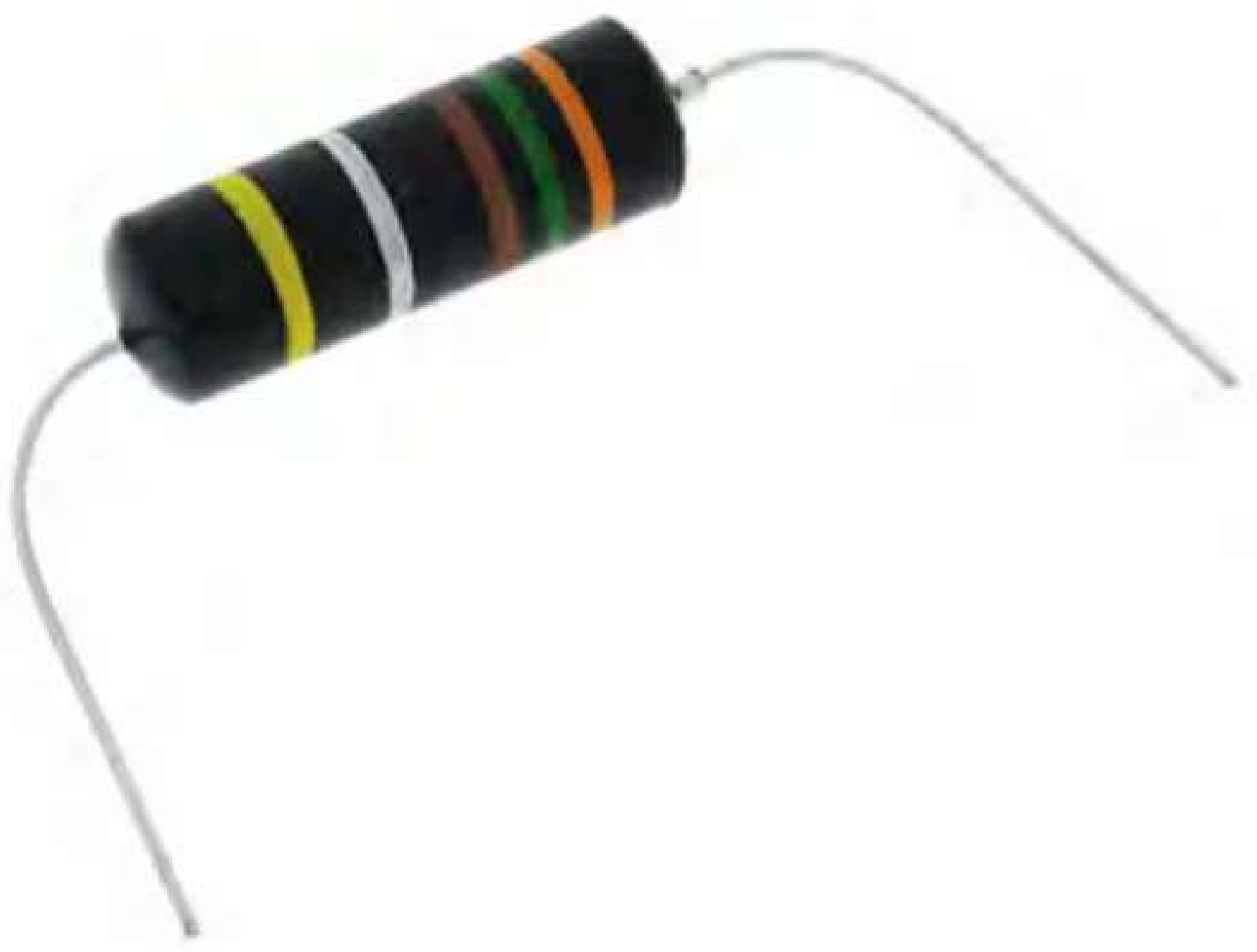

Fig.63: a Sprague "Bumble Bee" cap. This is a Thomann TAD Jupiter reproduction model for guitar restorers.

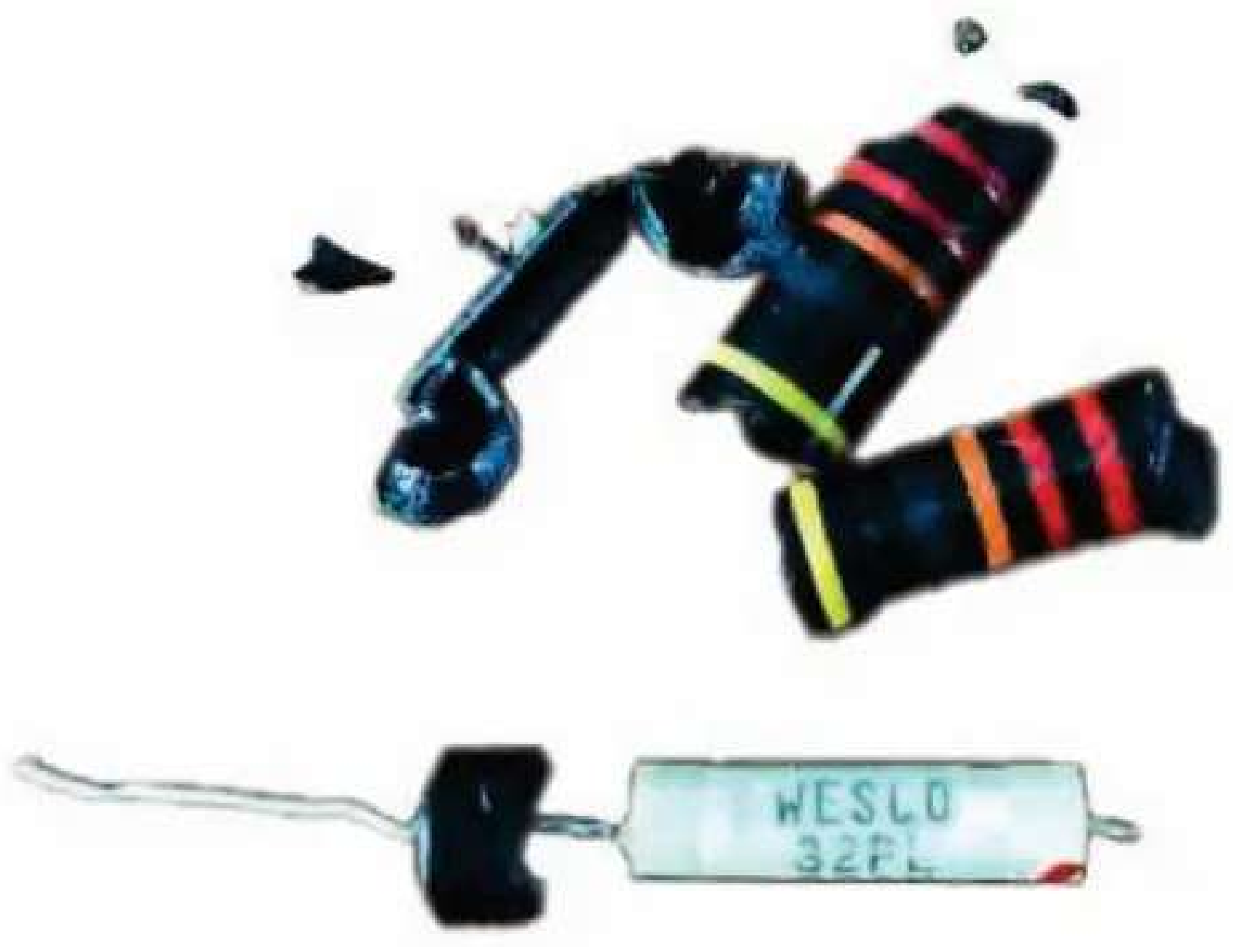

Fig.64: a fake "Bumble Bee" capacitor. Once opened, it reveals there was a cheap polyester capacitor inside it.

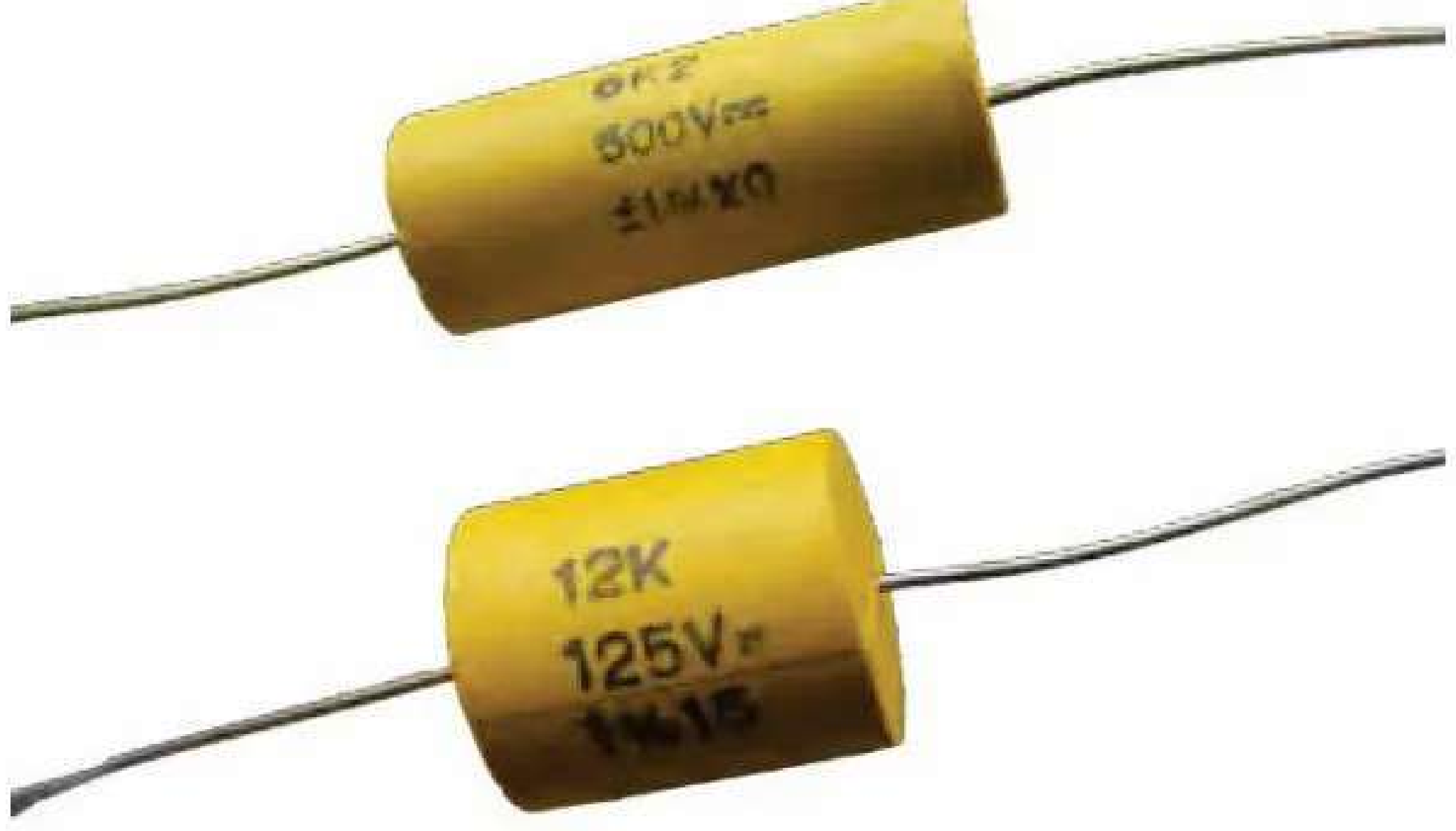

Fig.65: "Mini Poco" polystyrene capacitors. They are top-notch audio filter capacitors if rather large.

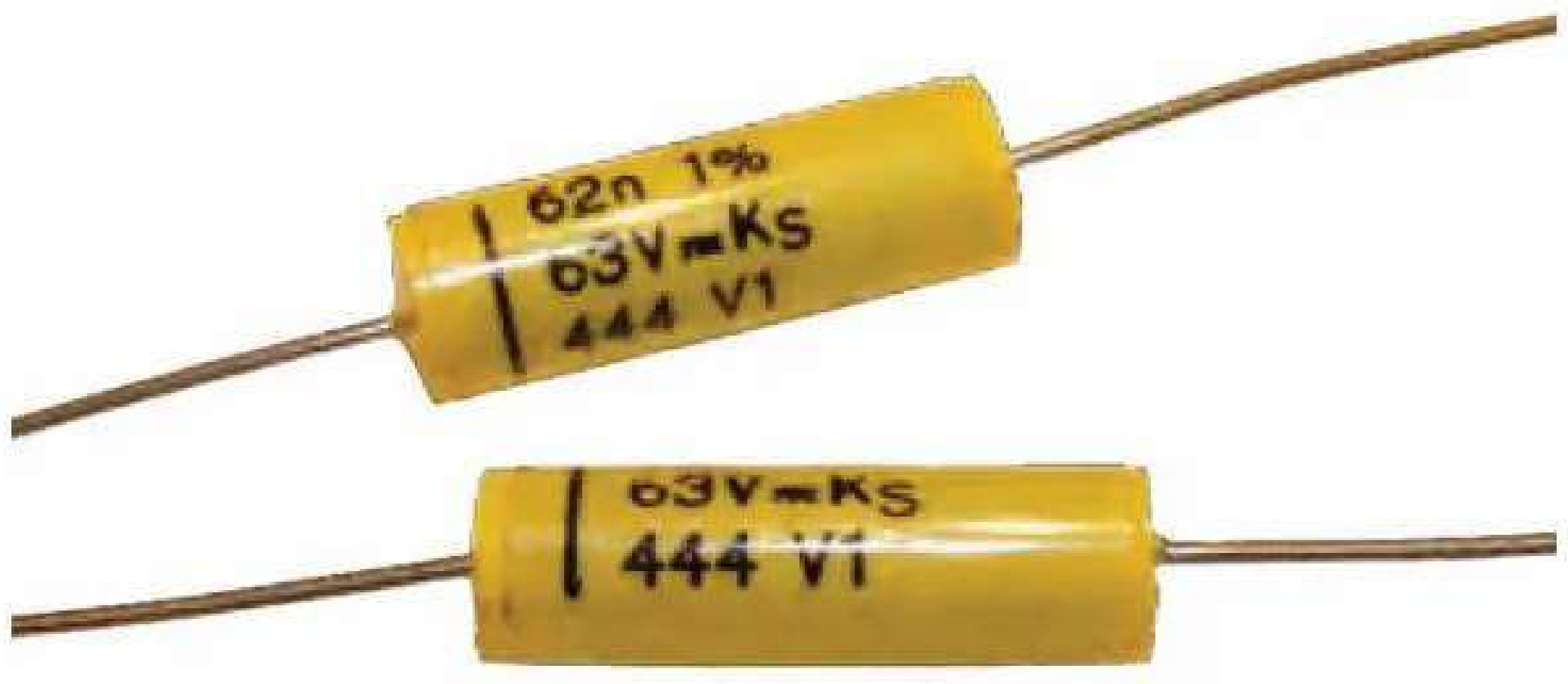

Fig.66: Philips 444-series polystyrene capacitors. Similar to the "Mini Poco" but with tougher epoxy encapsulation.

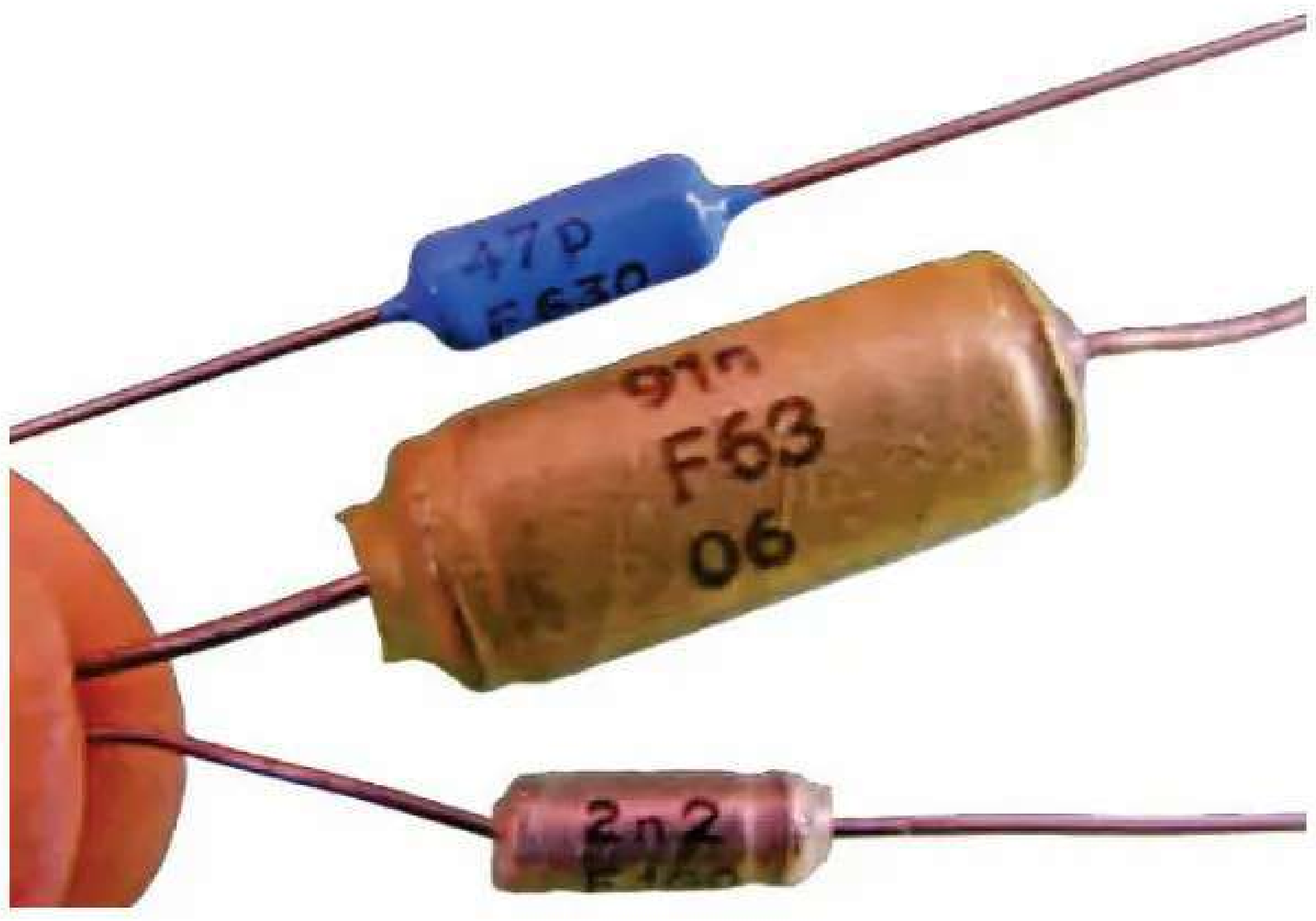

Fig.67: the final version of Philips' polystyrene capacitor. The blue cap is their polypropylene replacement.

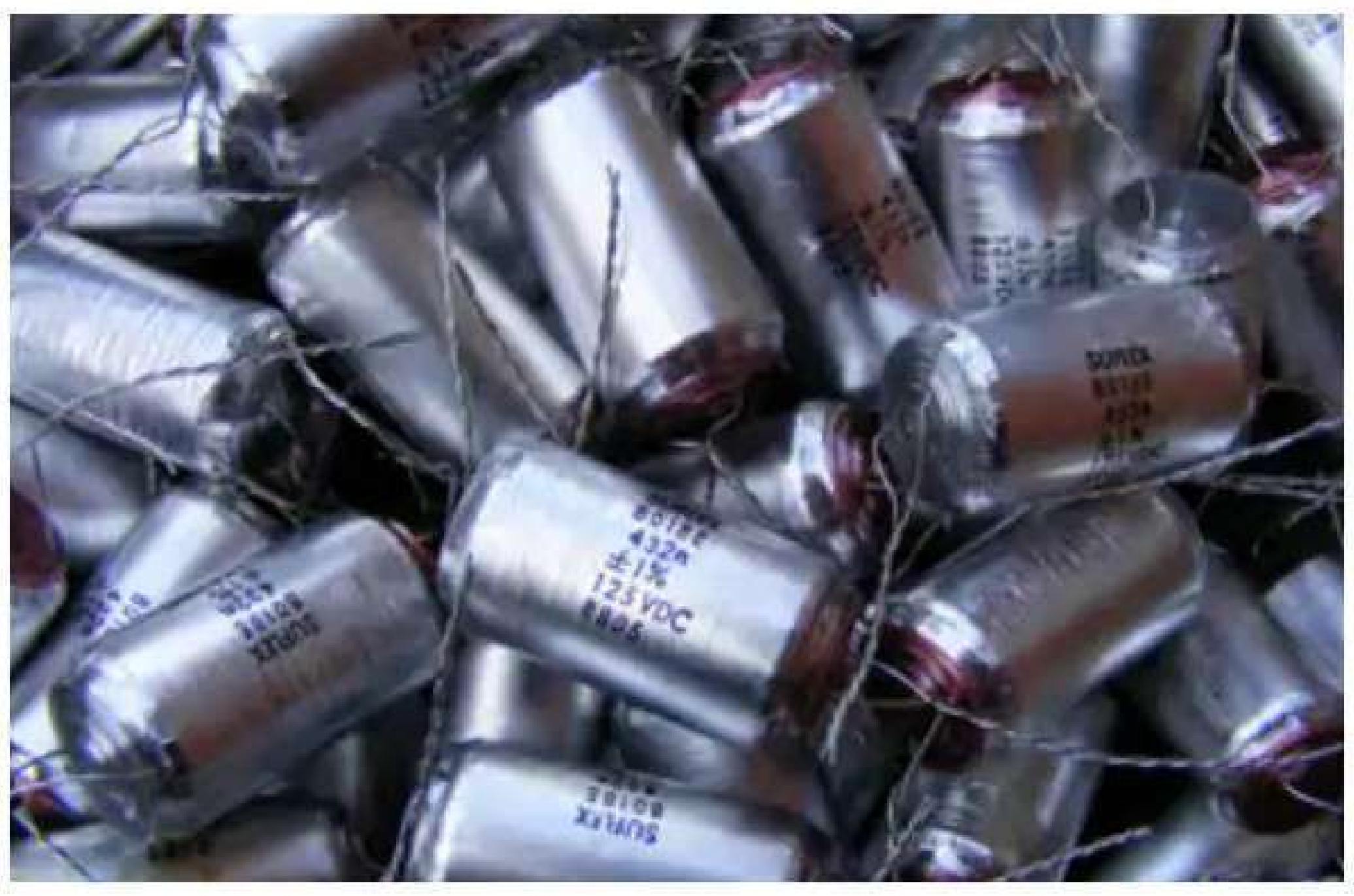

Fig.68: the biggest polystyrene I have seen, Suflex 432nF. The double leads give lower parasitic inductance.

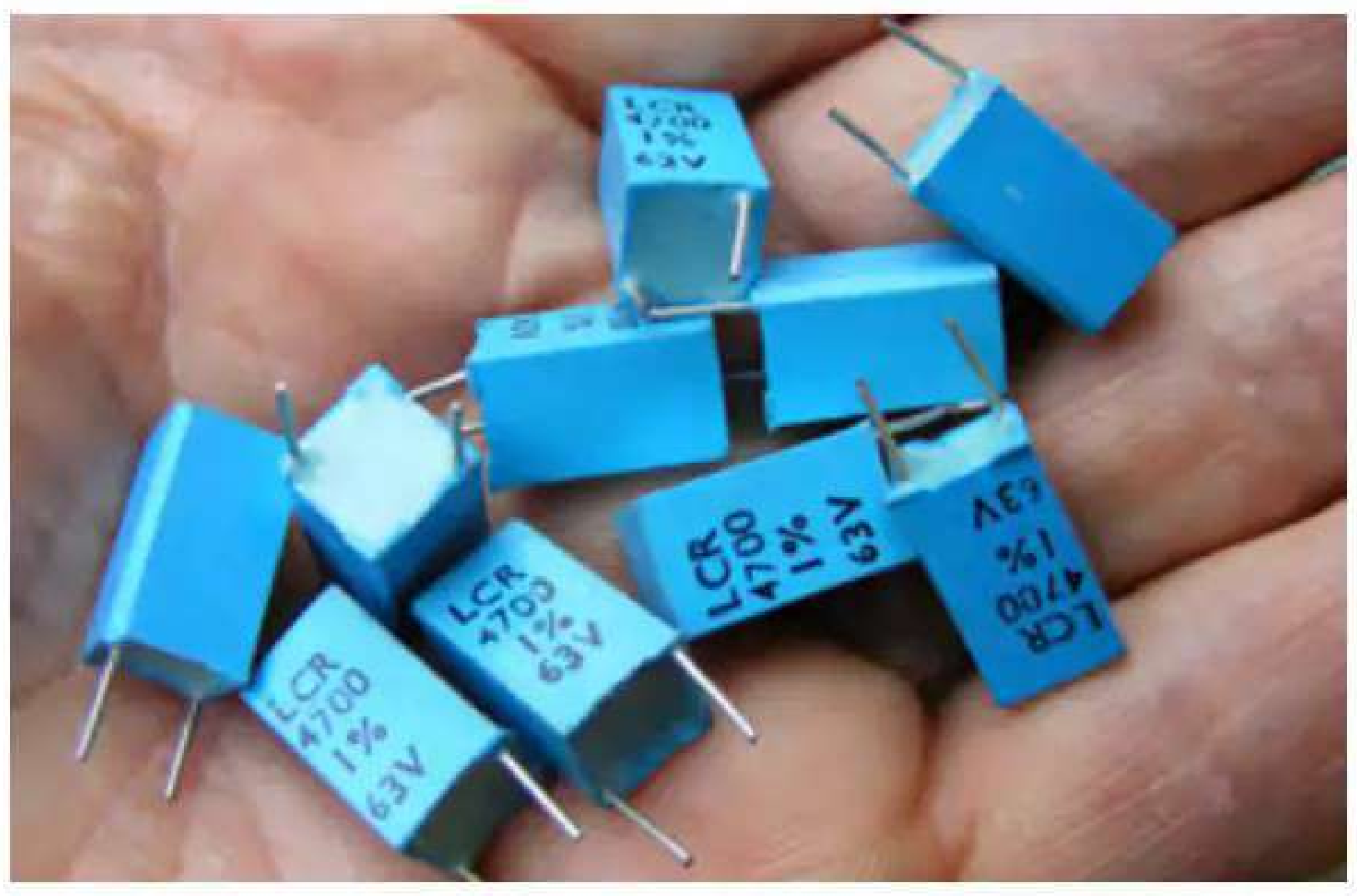

Fig.69: current polystyrene capacitors by LCR. Some versions now use tin foil.

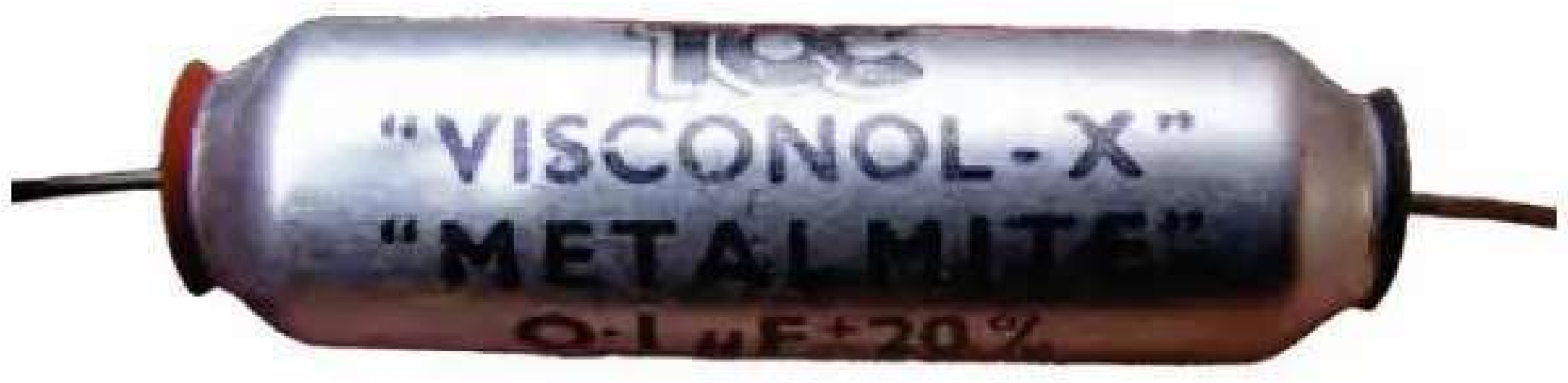

Fig.70: a TCC paper-in-oil capacitor from the Cold War.

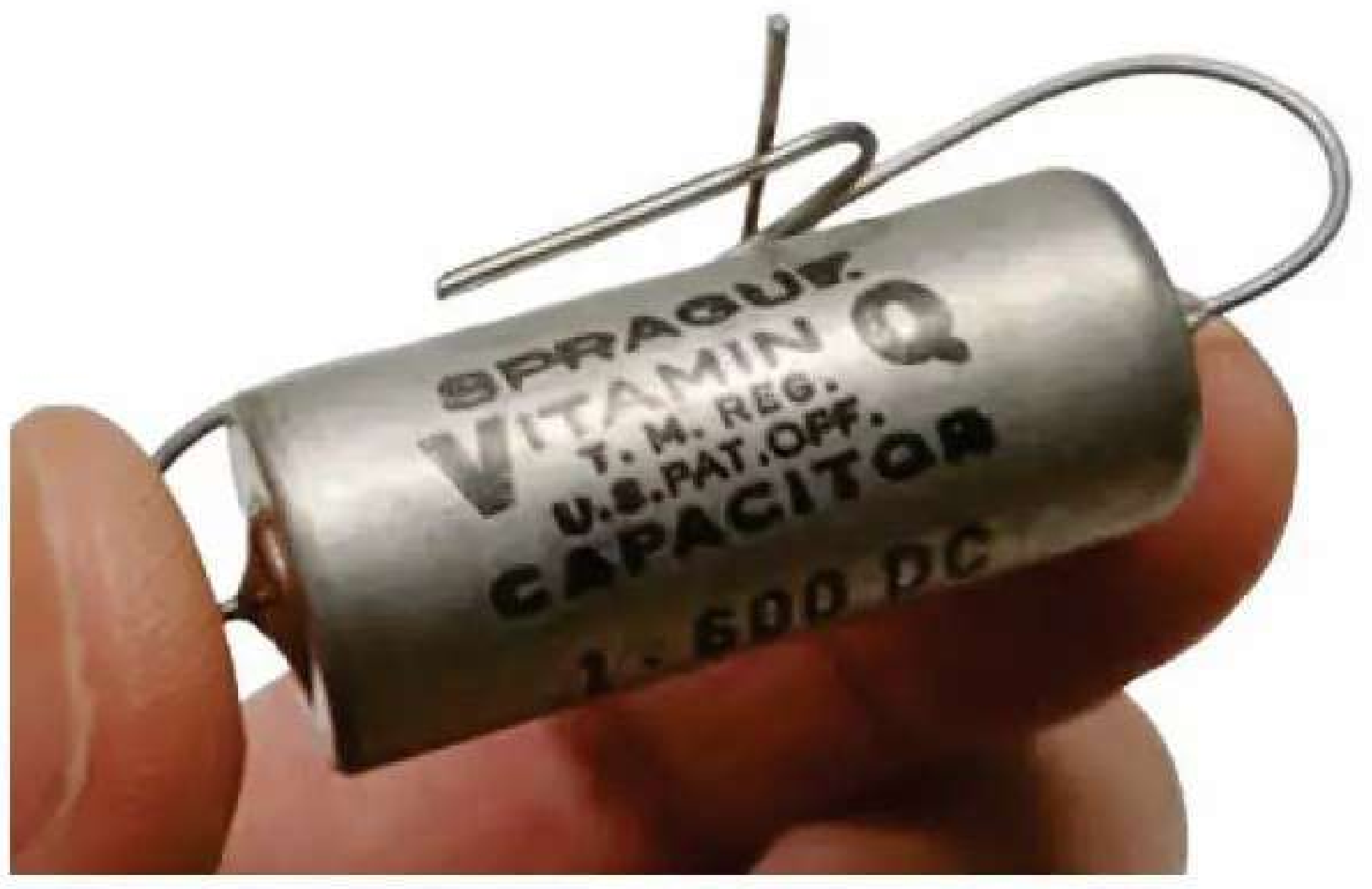

Fig.71: a "Vitamin Q" military paper-in-oil capacitor (from an old eBay photo).

replacement, the 444 series, shown in Fig.66.

The newer green version (Fig.67) was smaller and, because it had iron leads was okay with flow soldering. There is now a blue polypropylene replacement, which is almost as good.

Suflex HS polystyrene

These are very stable, low in cost and popular in RIAA networks in HiFi amplifiers. Their weakness is melting while soldering.

Suflex were the go-to choice if you wanted an odd custom value. I had an old lady neighbour who used to work at their Welsh plant. A sideline to her main assembly job was to inform MI5 where weird custom capacitors were going!

She gave me a few boxes before she died. There are some very odd values in there; 432nF ±1%, anyone (Fig.68)? I'll have to design a circuit especially for them.

The world's last polystyrene capacitors (Fig.69) are still made by LCR in Wales, who bought up Suflex's stock of film. They still make the HS series.

Sprague Vitamin Q

Paper-in-oil capacitors are highly favoured by valve amplifier makers. Like all wet capacitors, they have very low resonance. An old British paper-in-oil type by TCC Telegraph Condenser Company is shown in Fig.70. The most desirable type is the Sprague Vitamin Q (Fig.71) which, being hermetically sealed, doesn't suffer from electrical leakage.

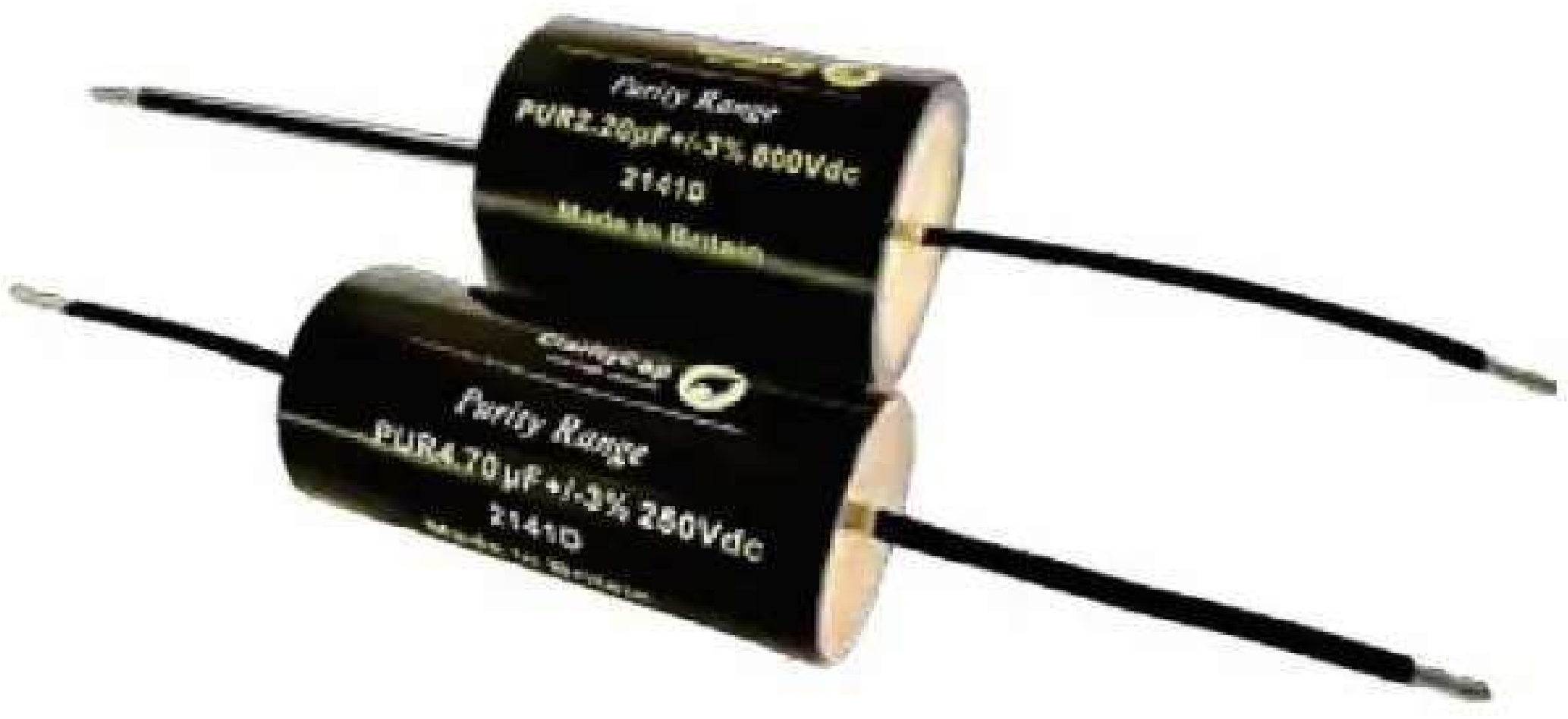

Fig.72: Clarity Caps – highly developed audio polypropylene (Charcroft).

Charcroft (ICW) Clarity Caps

I have to confess a vested interest here since I did some work for the owning company, Charcroft Electronics, who make these, and Salford University, who performed the research.

Clarity Caps are axial series-wound metallised polypropylene types with a mixture of polyurethane resin for damping and epoxy resin end-fill (Fig.72). This gives them the lowest mechanical resonance, around 15dB down compared to normal similar capacitors.

The end spray is zinc with a copper sponge matrix soldered on with thick copper leads. KEF use 5.7µF Clarity capacitors in their LX50 reference speakers.

One trick top HiFi companies use with specialised audio capacitors is to get their own branding printed on. Fig.73 shows a pair of branded slit-foil reservoir capacitors in a Cyrus amplifier.

So there you have it, just part of the mountain of tacit knowledge on the audio aspects of capacitors. I think I should write a book!

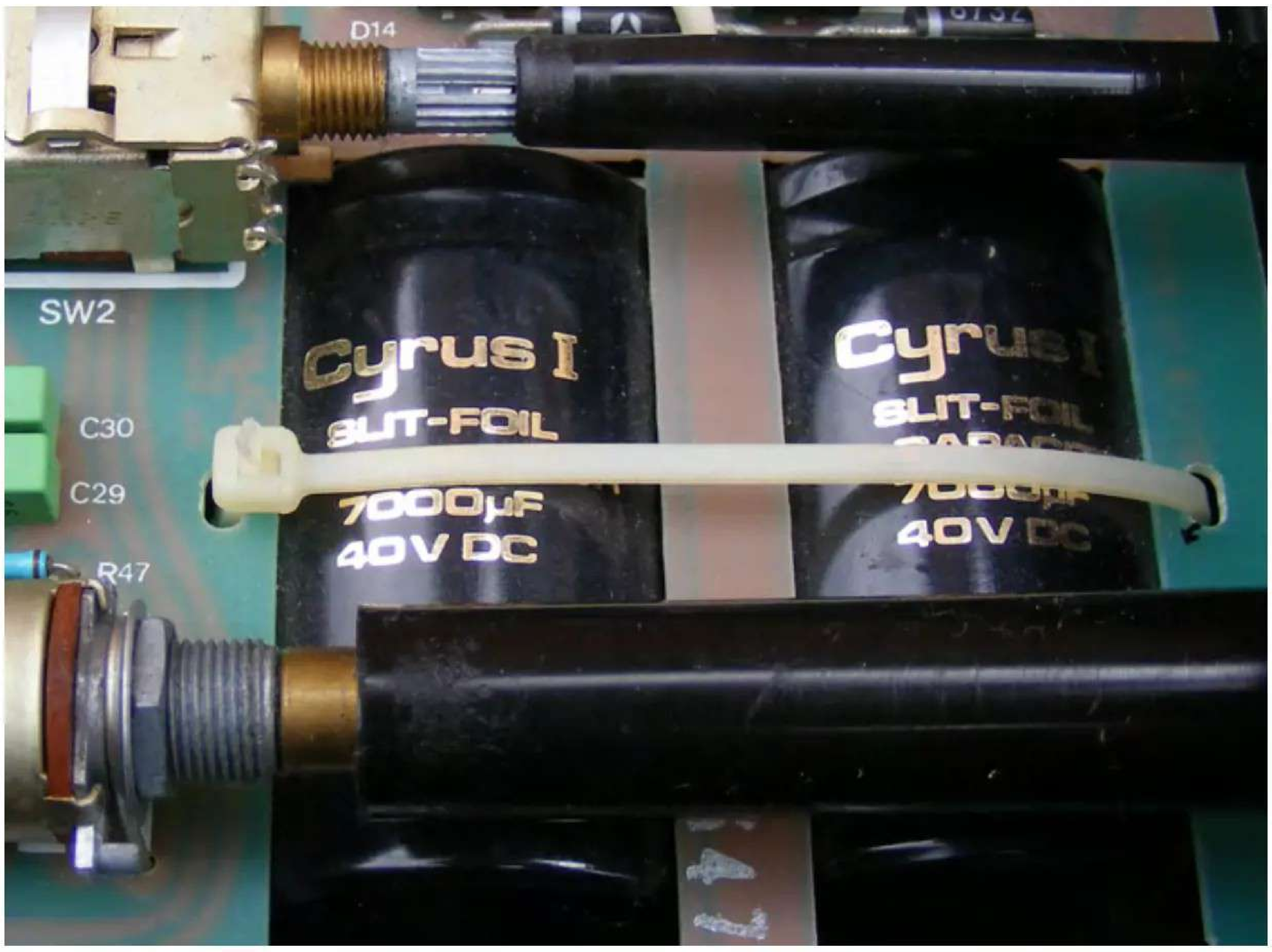

Fig.73: special slit-foil electrolytics in a Cyrus amplifier. Note the company's branding in gold. If you are paying for it, you should show it off!

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2SK2145-Y dual JFET	£0.80
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Note '10/63' denotes '10µF 63V'.

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5.6/63, 8.2/63, 10/63	£2.00
Mullard 'Mustard' C296 0.22/400	£2.00
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6.8/160V, 10/63	£2.00
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Cunning clones

Techno Talk

Max the Magnificent

Imagine that you could hold a conversation with holographic clones of your great-grandparents, asking them questions about their lives, loves, hopes and fears. It's too late for us, but this may be a possibility for our own great-grandchildren.

In my previous column, I introduced a company called Proto (see their website at protohologram.com). Its claim to fame is the ability to generate artificial intelligence (AI) based full-size 3D volumetric holograms of people.

They start by videoing a conversation with the person in question. Let's consider William Shatner—who famously played Captain James Tiberius Kirk in Star Trek: The Original Series (TOS)—as an example because they've already done him. The idea was to create a hologram of Bill that could attend shows and conferences and answer both Shatner- and Star Trek-related questions.

Proto's AI used this video to create what I think of as a 'voice clone'. The AI can subsequently use this clone to say anything it likes while sounding just like Bill. Their AI also creates what I think of as a 'physical clone' or a 'body clone', which it uses to generate the full-size 3D hologram. This hologram is generated on a frame-by-frame basis in real-time.

Any visual aspects of the hologram—lips, facial muscles, eye blinks, hand and body gestures—are synchronised with whatever it's saying.

Who are you?

To be capable of holding a Turing test level conversation, the AI powering the hologram needs what I think of as a 'knowledge clone' of the person in question. In the case of Bill, the AI can build this knowledge clone by interviewing him, starting with basic questions like, "What's your name?", "Where were you born?", "How old are you?" and "How do you pronounce 'sabotage'?".

Each question can open new avenues of conversational exploration, just like a chat between two people.

If they are lucky enough to get to meet Bill, many people want to pose Star Trek-related questions. Although Bill generally enjoys answering queries of this nature, he sometimes expresses frustration if he feels the questions are repetitive or overly focused on specific details. Unfortunately, fiddly details are what Star Trek fans like most.

The folks at Proto got around this by feeding the AI with every script from every episode, thereby allowing the hologram to answer any conceivable tribble-connected question without trouble (obviously everyone knows Season 2, Episode 15: "The Trouble with Tribbles").

Abby Normal

In the 1974 film Young Frankenstein, Frederick sends hunchback Igor to the local "Brain Depository" to steal the brain of a deceased "scientist and saint" called Hans Delbrück. Unfortunately, Igor accidentally drops the glass jar containing the brain in question, so he grabs another close by.

He tells Frederick the brain came from someone called "Abby Normal", but later admits the jar was, in fact, labelled "Abnormal".

With Proto's holograms, it's possible to swap one knowledge clone (the 'brain') for another. Suppose that instead of loading our Shatner AI with Star Trek scripts, we instead fed it with books on infectious diseases. This would enable it to hold forth on medical matters (just don't ask it about Tribbles).

Who said that?

The AIs powering Proto's holograms can typically recognise and speak over 120 languages (excluding Australian, of course).

Thus, if a Spanish fan approached our William hologram and asked, "¿Cuál es el verdadero problema con los tribbles?" (ie, "What's the real trouble with tribbles?"), the AI could (a) respond in Spanish complete with apical-alveolar trills (ie, the rolling of the Rs) and (b) the hologram's lips and other facial muscles would be perfectly synchronised with the words.

What's the point?

You might be thinking, "This is all very clever, but what's the point?" By this, you mean, "What could we use these AI holograms for?" To be honest, we are limited only by our imaginations.

Mainstream uses for this technology might include information booths at transportation hubs like airports (their human hosts always seem to be absent when I need them the most).

Or how about having one at the end of every aisle in a gigantic hardware store? These AIs can see and hear you, so you could show them something and ask, "What's the name of this and where can I find one?" With its subject matter expertise, the AI might respond, "That's called a ballcock valve, and you'll find them on the bottom shelf halfway down aisle 10 on the right."

A lot of people are embarrassed to talk to doctors about certain things. But experiments have shown that people are much more forthcoming with AIs, especially ones like Proto's holograms. And, just like a human doctor, the AI could be watching the patient, observing things like involuntary movements of the hands or rapid blinking or…

I can easily envisage these AI holograms in nursing facilities, especially those dealing with people suffering cognitive decline like dementia. Unlike human carers, an AI wouldn't mind answering the same question multiple times or listening to the same story over and over (and over) again.

The Life of (a boy called) Clive

I miss my dad. He passed away in January 2000. I would love to have access to an AI hologram of him—one that had spent weeks chatting to him and learning all about him. I know it wouldn't be like really having my dad present, but being able to hold a conversation with him (even in holographic form) would soothe my soul, as it were.

As I mentioned a couple of columns ago, I have almost finished writing a book called The Life of (a boy called) Clive. This describes my life growing up as a kid in Sheffield in the 1960s.

Suppose we fed the contents of this book to an AI hologram of me. The AI could also interview me to fill in additional details. Long after I'm gone, my great-great-grandchildren could chat with my holographic AI to at least get a glimpse as to what I was like. It would be a form of immortality. It's something to think about… PE

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Fuse 1A
SILICON CHIP
FAN SPEED CONTROLLER
(For shaded pole fans 80W maximum)
FAN
POWER

John Clarke's Mk2

Fan Speed Controller

On a hot night, a gentle cooling breeze from a fan can keep you cool and help you to sleep. This new Fan Speed Controller is an effective, noise-free, low-speed fan controller. It works with ceiling, pedestal and box fans.

Most fans include speed control, but many run too fast, even on their slowest setting, and can be pretty noisy. If you want to use the fan to keep cool while sleeping, you don't need a fast breeze but just gentle air movement. You also don't want the fan blades or the motor to make any noise that will keep you awake.

Whether a fan makes noise at a slow speed depends upon the type of speed control. Of the methods used for controlling fan speed, phase control causes the most motor noise. This type of control is where just a portion of the full mains sinewave is applied to the fan motor.

Because just a part of the mains waveform is applied, it produces a rapid change in voltage as the waveform is switched on and off. That can produce vibration in the motor windings and bearings, causing a buzzing sound.

Other fan speed controllers use a switch that selects from one of several different capacitors or inductors. While they don't generally make the fan motor noisy, they only provide a few fixed speeds and the lowest speed is usually not that slow.

Our Fan Speed Controller does not use phase control; instead, it introduces resistance in series with the fan motor to adjust the fan speed. The mains sinewave is simply reduced in voltage without changing the wave shape. Applying a sinusoidal voltage to the motor ensures the fan makes minimal noise. It also provides continuous adjustment from stopped to full speed or anywhere in between.

This does have the disadvantage that power is dissipated as heat. But considering that most fans will draw a maximum of 60W at full speed and less as speed is reduced, the heat produced is modest and can be dissipated by the aluminium diecast box, which acts as a heatsink.

We don't need to dissipate anywhere near 60W because, at full speed, the dissipation in the controller is relatively low since the resistance of the controller is low. At lower speeds, where the controller resistance is higher, dissipation increases. But because the motor is running slower, the overall power drawn by the fan is much less than at full speed.

An over-temperature thermostat will switch the power off should there be excess heat buildup. This precaution prevents the speed controller from overheating and possibly causing skin burns if touched.

For the resistance element, we use a Mosfet with a drain-to-source resistance that can be controlled by adjusting the gate voltage. The Mosfet can behave like a very low resistance for full-speed operation or a higher resistance under partial conduction for slower speeds.

A single Mosfet cannot directly control the mains AC voltage. While it operates as a resistance element when the current flows in one direction, in the other direction, it is shunted by an intrinsic diode that's part of the Mosfet structure.

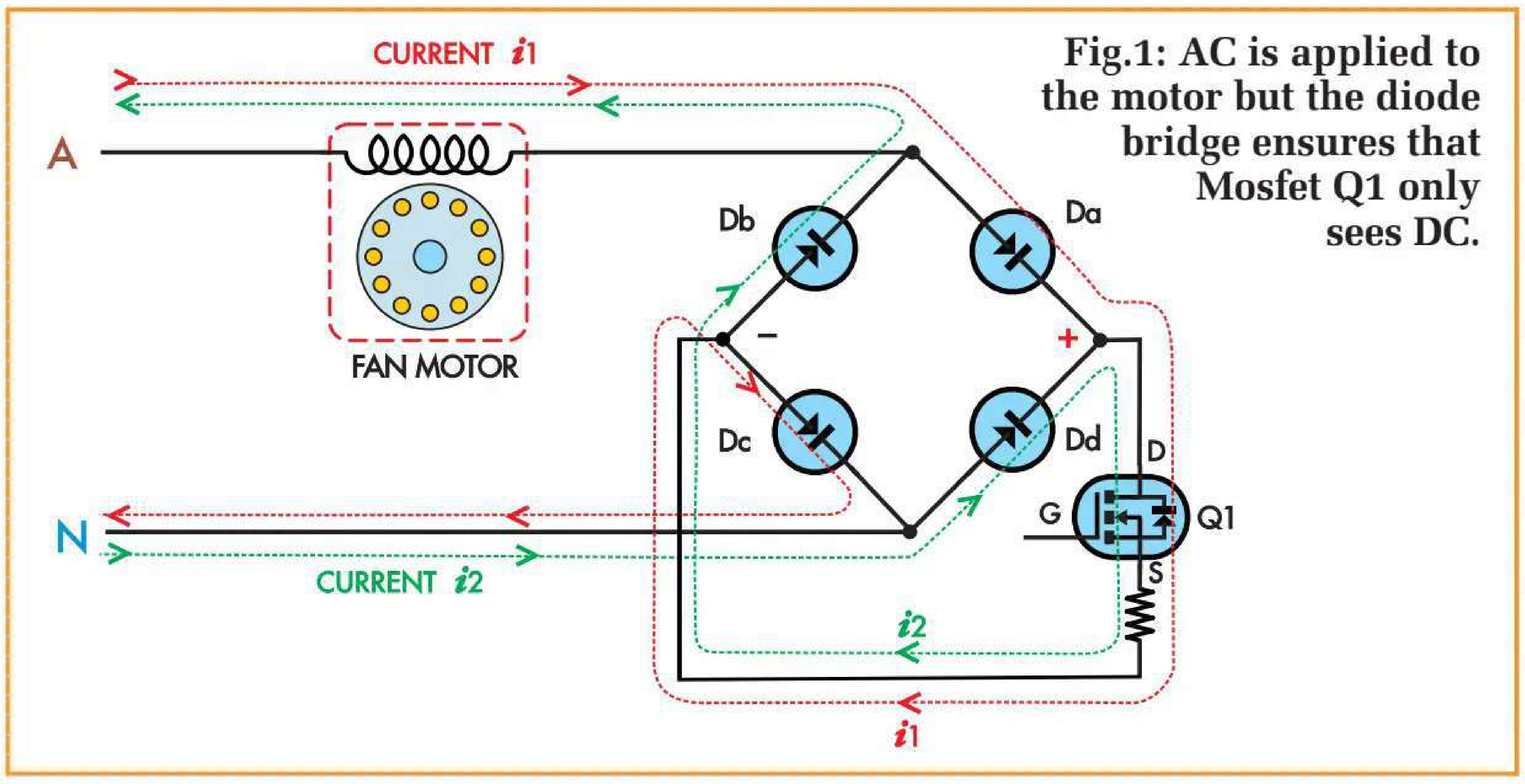
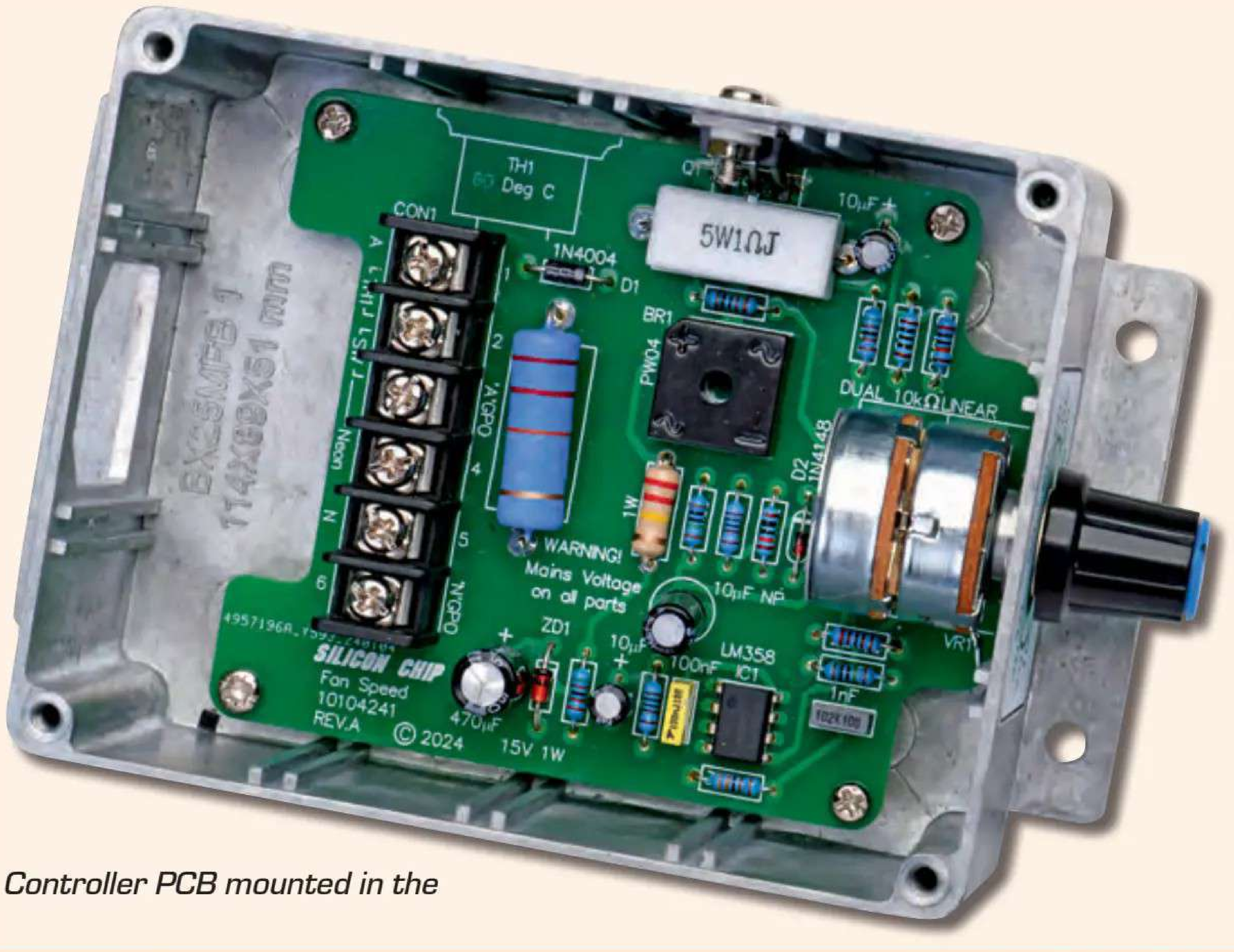


Fig.1: AC is applied to the motor but the diode bridge ensures that Mosfet Q1 only sees DC.

Fan Speed Controller, Mk2 Features & Specifications

- Quiet fan speed control
- Suitable for 230V AC shaded-pole fan motors
- Full control of motor speed from stopped to maximum
- Over-current limiting
- Over-temperature cutout
- Fuse protection against faults
- Rugged aluminium case
- Fan power: 80W maximum
- Fuse: 1A, 230V AC
- Current limiting: 235mA at low speed, up to 940mA at high speed
- Over-temperature cutout: triggers with case at 50°C (resumes at 45°C)



This photo shows the completed Fan Speed Controller PCB mounted in the case without any of the wiring.

To prevent reverse current flow through the Mosfet, the Mosfet is placed within a full wave bridge rectifier. That way, it only handles current in one direction, but an alternating current (and voltage) is still applied to the fan.

Fig.1 shows the general arrangement. The Mosfet (Q1) is between the positive and negative terminals of the bridge rectifier. When the mains Live voltage is more positive than the Neutral, current (i1) flows from Live through the motor, diode D_a, Mosfet Q1, then diode D_c to Neutral.

When the Live is more negative than Neutral, current (i_2) flows from Neutral through diode D_d, Mosfet Q1, diode D_b and the fan motor to Live. In both cases, the current through Mosfet Q1 is always from its drain to its source and never in the reverse direction, so the current never flows through the body diode.

Full circuit description

The circuit for the Fan Speed Controller is shown in Fig.2. It comprises just one IC, several diodes, the high voltage Mosfet, Q1, plus some resistors and capacitors. Power for the circuit is derived directly from the 230V AC mains.

The entire circuit floats at mains potential, including circuit ground, which is not connected to mains Earth.

The critical part of the circuit comprises potentiometer VR1b, op amp IC1a and Mosfet Q1. This part of the circuit allows the user to adjust the average voltage across Mosfet Q1 using potentiometer VR1b. As VR1b is rotated clockwise, the voltage applied to pin 2 of IC1a reduces. IC1a reacts by increasing the gate voltage of Mosfet Q1 to reduce the average voltage across its channel.

That might seem backward, rotating clockwise to reduce the voltage. However, Q1 is in series with the fan motor, so the fan gets more voltage when the voltage between Q1's drain and source is lower.

So when VR1b is fully anticlockwise, the average voltage across Q1 is at a maximum, and the applied voltage to the fan is at a minimum. As VR1b is rotated clockwise, the voltage across Q1 decreases, and the voltage applied to the fan increases, allowing it to speed up.

At the same time, IC1b monitors the current through Q1 and provides current limiting to prevent excessive current flow that could overheat and damage Q1. That usually should not happen, but it depends on what is plugged into the outlet. Perhaps someone will plug in a fan that's too large or a different load, in which case IC1b will activate to protect Q1.

In more detail

Op amp IC1a, which drives the gate of Mosfet Q1, is connected in a feedback control loop that monitors a divided version of the voltage across Q1's channel (drain to source) and the voltage from the wiper of speed potentiometer VR1b. IC1a adjusts its output voltage so the divided Mosfet channel voltage matches that set by the speed potentiometer.

The divider is formed by a 220kΩ 1W resistor and a 5.1kΩ ¼W resistor. The voltage from this divider is filtered with a 10μF capacitor, providing a DC voltage proportional to the average of the full-wave rectified voltage.

The resistive divider is there to produce a voltage suitable for monitoring by IC1a. When monitoring up to 230V AC (325V DC peak), the divider output is around 7.4V peak that averages to 4.7VDC after filtering. This average voltage is 63.7% of the waveform peak voltage and well within the input range for IC1a when powered from a 15V supply.

As the resistance of Q1 decreases and the fan speeds up, there is more voltage across the fan motor and less voltage across the Mosfet. The voltage from the divider therefore also reduces.

The Mosfet source also has a 1Ω series resistor that connects it to circuit ground for current monitoring. This increases the voltage applied to the divider by about 1V, depending on the fan motor current, but this does not affect the output from the voltage divider much. That's because 1V is a small fraction of the hundreds of volts that can be across the Mosfet.

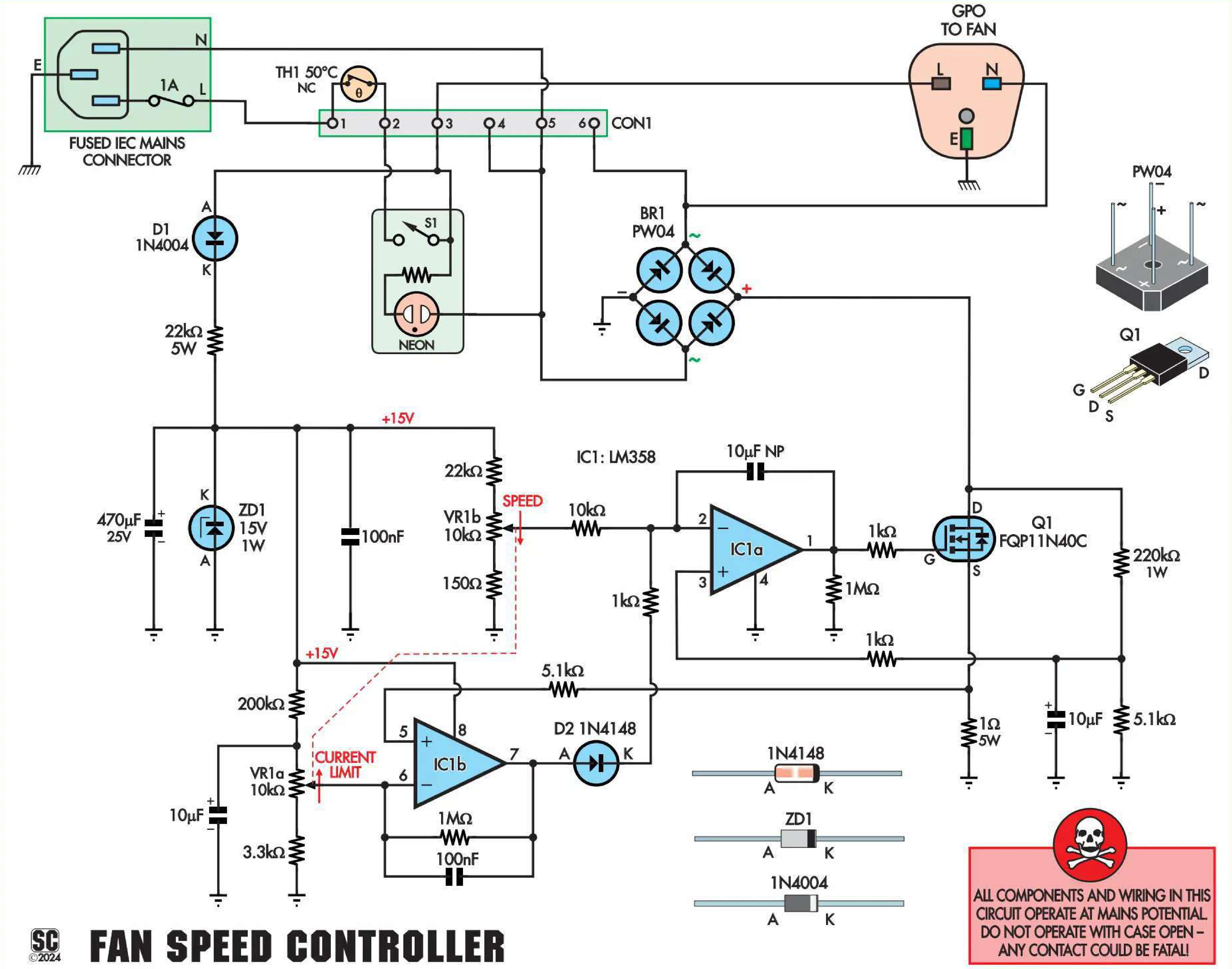

GPO
TO FAN
L
N
E
FUSED IEC MAINS
CONNECTOR
1A
TH1 50°C
NC
θ
1
2
3
4
5
6
CON1
PW04
~
+
Q1
G
D
S
D1
1N4004
A
K
S1
NEON
BR1
PW04
22kΩ
5W
+15V
IC1: LM358
10µF NP
22kΩ
SPEED
VR1b
10kΩ
10kΩ
150Ω
470µF
25V
ZD1
15V
1W
100nF
IC1a
1kΩ
1MΩ
1kΩ
Q1
FQP11N40C
220kΩ
1W
5.1kΩ
+15V
200kΩ
CURRENT
LIMIT
VR1a
10kΩ
10µF
3.3kΩ
IC1b
D2 1N4148
1MΩ
100nF
1Ω
5W
10µF
5.1kΩ
1N4148
ZD1
1N4004
ALL COMPONENTS AND WIRING IN THIS CIRCUIT OPERATE AT MAINS POTENTIAL. DO NOT OPERATE WITH CASE OPEN – ANY CONTACT COULD BE FATAL!
SC
©2024
FAN SPEED CONTROLLER

Fig.2: the circuit diagram for the Fan Speed Controller Mk2. Op amp IC1a controls the resistance of Mosfet Q1 to regulate the fan speed while op amp IC1b prevents the fan from drawing too much current.

Potentiometer VR1b is used to set the fan speed. It connects in series between a 22kΩ resistor from the +15V supply and a 150Ω resistor to the 0V supply. With this resistor string, the voltage range for VR1a's wiper is 5V to 0.07V. The lower voltage for VR1b is deliberately made to be slightly above 0V as IC1b would oscillate if it were set to 0V.

Another reason for keeping the lower limit at 70mV is to avoid the Mosfet operating outside its safe operating area, but more about that later.

If VR1b is set to produce 2V DC at its wiper, IC1a adjusts its drive to the gate of Q1 so that the voltage monitored at the resistive divider junction is also 2V DC. 2V on the divider means that there is 88V average across Q1, equivalent to 97.5V RMS. If the mains voltage is 230V AC, the voltage across the fan is 230V minus 97.5V or 132.5V RMS.

The feedback control ensures that the voltage across the Mosfet is strictly maintained to prevent changes in the motor speed. Without the feedback control, just applying a fixed voltage to the gate of Q1, the fan would slow quite markedly as the Mosfet heats up. That's because the Mosfet drain-to-source resistance increases with temperature.

Apart from adjusting the speed control (VR1b), the only other factor that can alter the fan speed is if the mains voltage changes. Typically, the mains voltage is reasonably constant, fluctuating by less than 5%.

Current limiting

Current limiting for this circuit is necessary since we are operating the Mosfet in a linear mode for speed control. Linear operation has the Mosfet operating in a region of partial conduction where it is neither fully conducting (with minimal on-resistance) nor fully off. This differs from a switching circuit where the Mosfet is either fully on or off.

Linear operation sees the Mosfet dissipating significant amounts of power, so the Mosfet must be kept within the safe region of its drain current (I_D) versus drain to source voltage (V_{DS}) over the entire voltage range. The manufacturer's safe operating area (SOA) graph for the Mosfet shows the region of operation.

Fig.3 shows the DC SOA curves for three different Mosfets that can be used in this circuit. SOA graphs also show the pulsed region of operation, but since we are not switching the Mosfet on and off, we have only included the DC SOA curves. These keep the Mosfet semiconductor junction below 150°C.

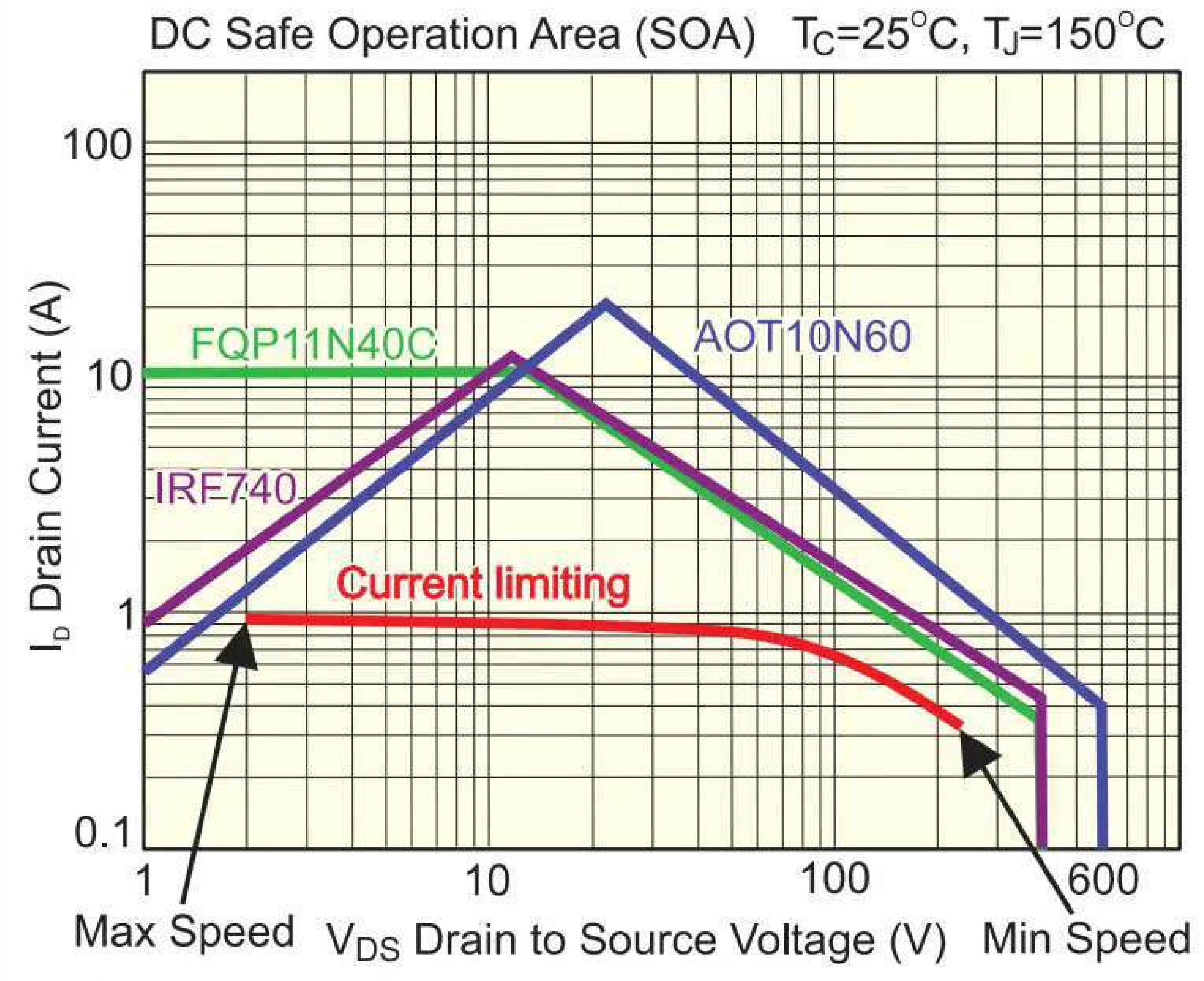


Fig.3: the DC SOA (safe operating area) for three different Mosfets. The current limiting curve is well within all three. Two of the curves are limited by the minimum Mosfet on-resistance at lower voltages, so even if the red line was extended to lower V_{DS} values, it couldn't cross them.

For each Mosfet to be used safely, we need to keep the curve in the operating region below the DC SOA curve. If the Mosfet is operated above the curve, it will likely fail due to melting (maybe not immediately, but eventually).

The red line indicates our circuit's current limit to safeguard the Mosfet. We restrict the maximum current to around 1A up to about 20V V_{DS}. Up to 20V, the fan will run fast. The V_{DS} will be higher at lower fan speed settings, so we reduce the current limit to prevent it from encroaching on the SOA curve. For the slowest speeds and highest V_{DS}, the current is limited to around 230mA.

That does not mean the Mosfet will be operating near this curve. It is just an overload threshold where the Mosfet is protected from damage, should conditions cause the Mosfet operating point to otherwise go above the current limit curve.

IC1b monitors the voltage across the 1Ω 5W resistor in series with Q1 for current limiting. This resistor converts the fan current to a voltage; eg, at 1A, it has 1V across it.

IC1b is connected as an amplifier with a level shift due to VR1a. As the voltage across the 1Ω resistor exceeds the voltage at the wiper of VR1a, IC1b's output goes high and drives the pin 2 input of IC1a high via diode D1 and the 1kΩ series resistor. This overrides the motor speed setting of VR1b, slowing the fan speed to reduce the current.

If the voltage across the 1Ω resistor is less than the voltage set at the wiper of VR1a, IC1b's output is low and has no effect on IC1a, as D2 is reverse-biased.

VR1a is connected across the 15V supply similarly to VR1b, but the padder resistors have different values. The 200kΩ and 3.3kΩ resistors set VR1a's wiper range to 235-940mV. VR1a and VR1b share the same shaft, so adjusting the fan speed will automatically adjust the current limit.

Note that VR1b's wiper produces a lower voltage as the potentiometer is rotated clockwise, while VR1a's wiper voltage increases as it is rotated clockwise. That's so that the current limit is higher for faster fan speeds.

Power supply

Mains power is applied to the controller via fuse F1, which is within the IEC input connector. This protects the circuit against excessive current flow should a fault occur, such as a broken wire short-circuiting against the enclosure. Both power switch S1 and over-temperature switch TH1 must be closed for the Live mains conductor to be connected.

Switch S1 includes a Neon indicator that lights when the unit is switched on.

TH1 monitors the temperature of the fan speed controller enclosure and switches off power if it reaches 50°C. It will reconnect power once the temperature drops to 45°C. This 5°C temperature hysteresis prevents the controller from switching on and off rapidly since it will take some time to cool by 5°C. The Neon indicator within S1 will be unlit whenever TH1 is open.

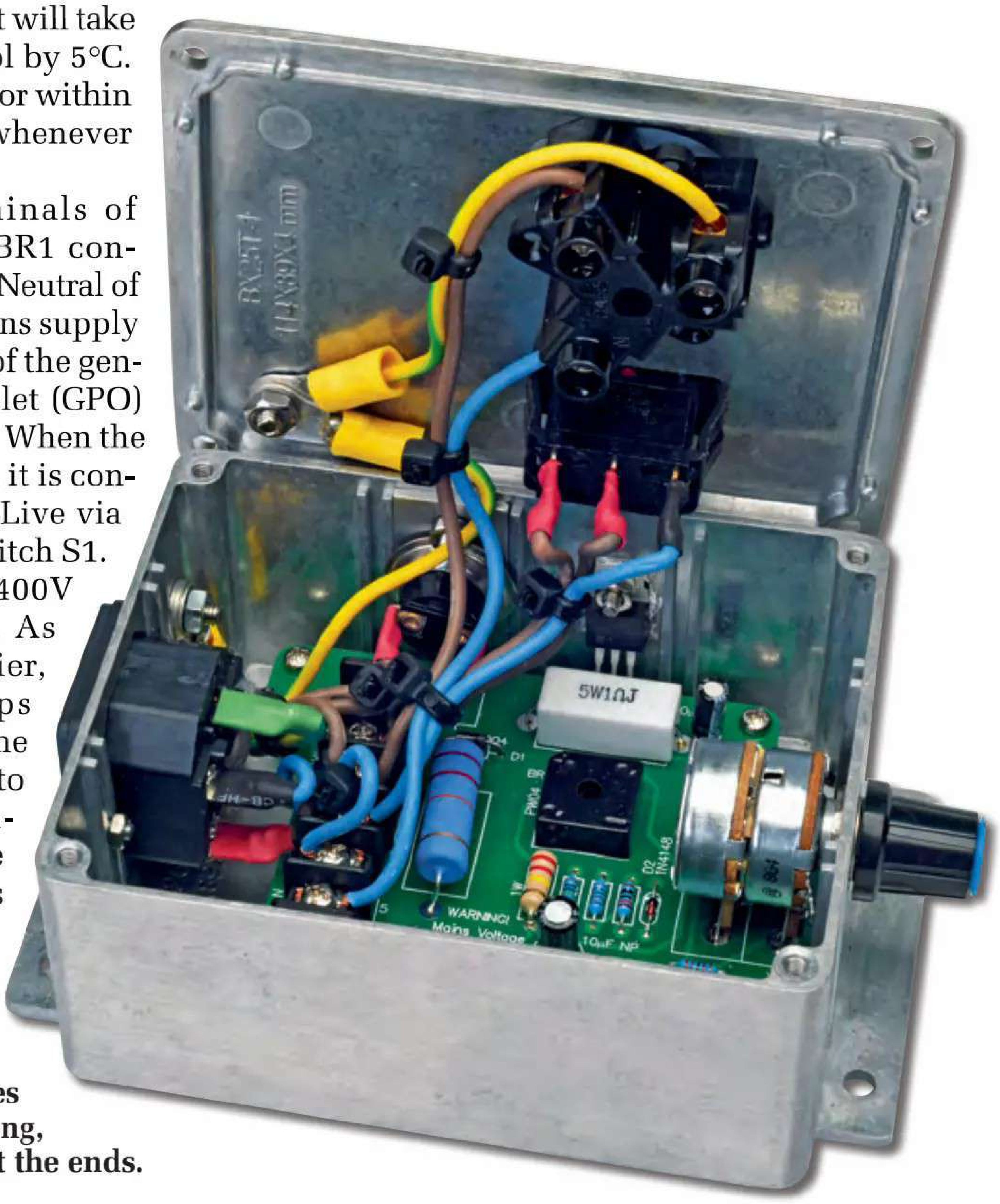
The AC terminals of bridge rectifier BR1 connect between the Neutral of the incoming mains supply and the Neutral of the general purpose outlet (GPO) for the fan motor. When the fan is connected, it is connected to mains Live via the GPO from switch S1.

BR1 is a 6A, 400V bridge rectifier. As mentioned earlier, the bridge keeps the polarity of the voltage applied to the Mosfet consistent while the fan motor receives AC.

A 15V supply to power the remainder of the circuit is obtained using a 22kΩ dropping resistor via diode D1 directly from the 230V AC mains switched Live. A 470μF capacitor filters the rectified waveform to produce a smoothed DC voltage clamped to 15V by zener diode ZD1. This 15V supply powers dual op amp IC1, Mosfet Q1 and the associated diodes, resistors and capacitors.

Using an X2 capacitor instead of a 22kΩ resistor would be slightly more efficient, like the previous design from the May 2015 issue.

However, the capacitor is somewhat



Make sure to use plenty of cable ties to secure the wiring, and heatshrink at the ends.

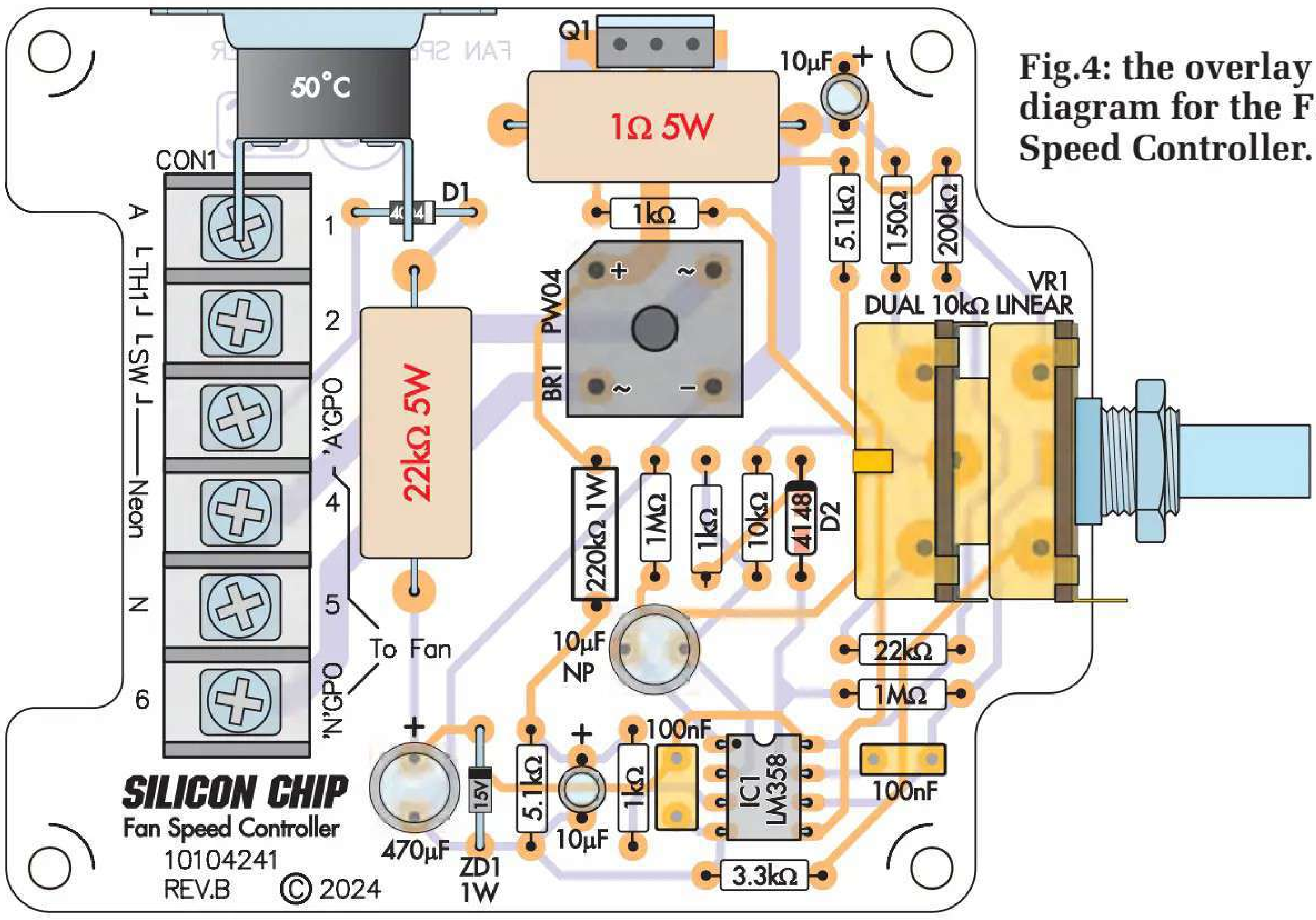

Q1
50°C
1Ω 5W
10µF +
CON1
D1
1kΩ
5.1kΩ
150Ω
200kΩ
A
L TH1
L SW
Neon
N
1
2
3
4
5
6
'A'GPO
'N'GPO
22kΩ 5W
PW04
BR1
VR1
DUAL 10kΩ LINEAR
220kΩ 1W
1MΩ
1kΩ
10kΩ
4148
D2
To Fan
10µF NP
22kΩ
1MΩ
100nF
IC1 LM358
100nF
+
470µF
15V
ZD1 1W
5.1kΩ
10µF
1kΩ
3.3kΩ
SILICON CHIP
Fan Speed Controller
10104241
REV.B
© 2024

Fig.4: the overlay diagram for the Fan Speed Controller.

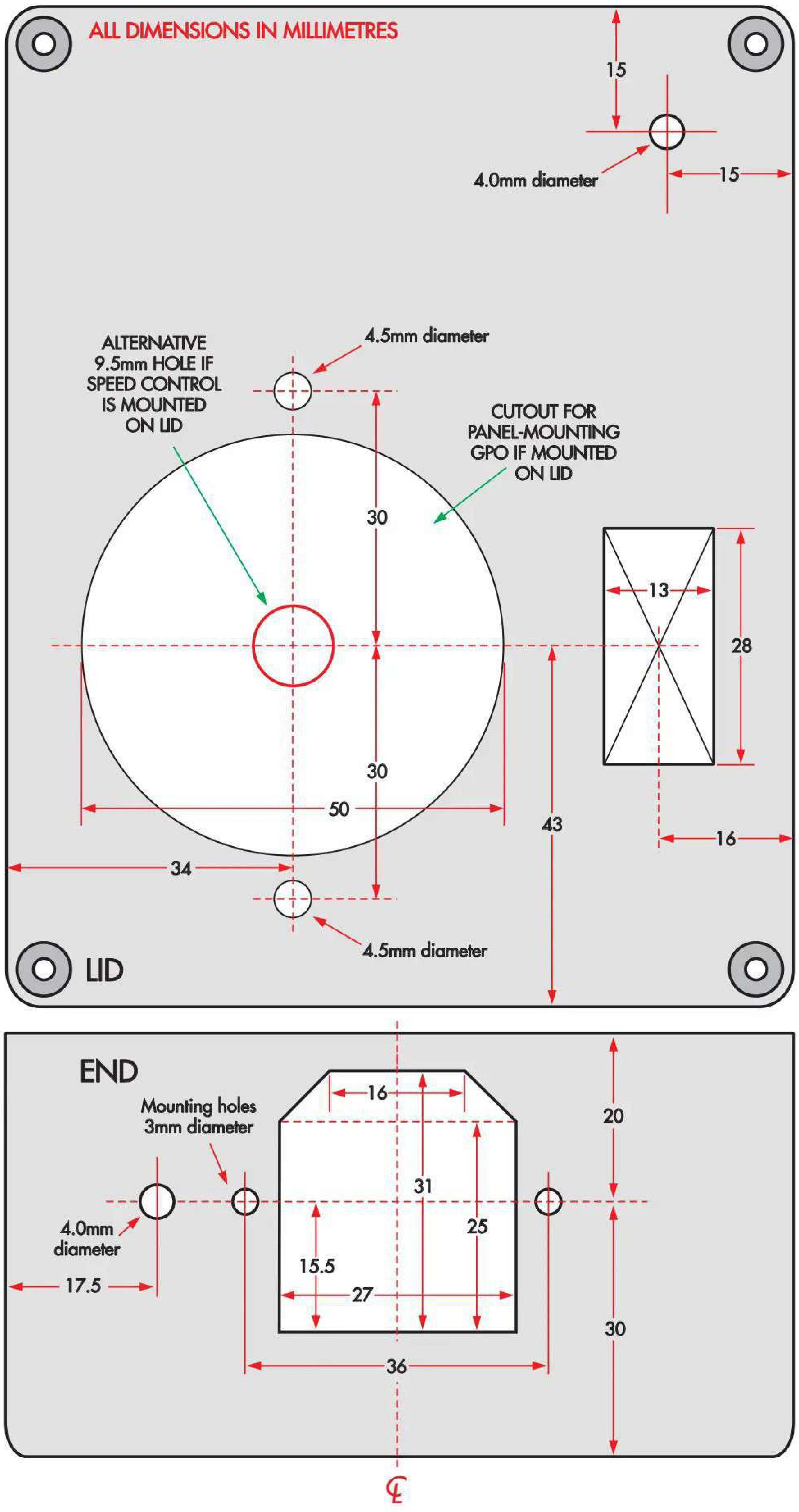

ALL DIMENSIONS IN MILLIMETRES
15
15
4.0mm diameter
ALTERNATIVE 9.5mm HOLE IF SPEED CONTROL IS MOUNTED ON LID
4.5mm diameter
CUTOUT FOR PANEL-MOUNTING GPO IF MOUNTED ON LID
30
30
13
28
50
43
16
34
4.5mm diameter
LID
END
Mounting holes 3mm diameter
16
20
31
25
4.0mm diameter
15.5
27
17.5
36
30
℄

Fig.5: the cutting and drilling guide for the diecast aluminium case. Depending on the application, the potentiometer can project from the side of the case or the lid, so read the text before making any holes. The red circle shows the hole for the shaft when mounting the pot on the lid (which is only possible if the GPO is not used).

While the prototype has an Australian mains outlet, this cut-out has been adjusted to suit the UK socket specified in the parts list. Both countries use a nominal 230V AC so the socket is the only change required.

expensive and bulky, and requires other support components like a second bridge rectifier. We decided it was not worth the size, expense or complexity for a slight increase in efficiency.

Enclosure & mounting options

Three different diecast aluminium enclosures can be used to house the Fan Speed Controller: an IP65/IP66 diecast box measuring 115 × 90 × 55mm (Gainta BS25MF, available from TME) or an economy diecast box measuring 119 × 94 × 57mm (Bud Industries CU-234), which is stocked by Farnell, DigiKey and Mouser.

The PCB is shaped so that it fits within the contours of the Gainta case, allowing it to be mounted horizontally on the enclosure's integral lands. For the Bud Industries enclosure, there are minimal internal contours to avoid but also no integral PCB-mounting lands, so the PCB needs to mount using four 9mm Nylon standoffs, attached via holes drilled in the base.

The Fan Speed Controller can be built as a standalone controller that plugs into a mains socket for power and has a general purpose outlet (GPO) that the fan plugs into. This version is suitable for pedestal and box fans.

For ceiling fans, the Fan Speed Controller can be built to intercept the fan wiring at the wall switch. In this case, it will need to be installed by a licensed electrician. The speed control adjustment potentiometer can be placed at one end of the enclosure, like the standalone version, or on the lid, which may be more convenient if the enclosure is wall-mounted.

Construction

The Fan Speed Controller is built on a PCB coded 10104241 that measures 94 × 79mm. To assemble it, follow the overlay diagram, Fig.4. Begin by soldering in the resistors, using the table for the colour codes in the parts list, but leave the 5W resistors off for the moment. Diodes D1, D2 and ZD1 can be fitted next, taking care to orientate them correctly (and don't get the three different types mixed up).

You can use an IC socket for IC1, or it can be directly soldered in. The latter should give better long-term reliability. Either way, be sure to install the socket and the IC correctly, with the notch facing the direction shown on the overlay. Then mount the two 5W resistors, slightly raised from the PCB surface, to aid in cooling.

Install the capacitors next. The 100nF capacitor may be labelled as 104. The electrolytic capacitors have their value directly marked and must be orientated correctly, with the longer leads through the holes marked with a + symbol. However, the larger 10µF capacitor is non-polarised (NP) and can be mounted either way around.

Fit diode bridge BR1 now, taking care that the cut corner is towards the top left of the board and placed adjacent to the + symbol.

Before installing VR1, its shaft may need to be cut to length to suit its knob. Do not install the potentiometer on the board if it is to be mounted on the lid. The six-way screw terminal strip (CON1) can be fitted now.

Q1 is mounted by kinking the outer two leads outward so that they will fit into the more widely spaced holes in the PCB. This wider spacing provides a 2.54mm clearance between the Q1 mounting pads and prevents possible arcing between the leads with peak voltages approaching 400V. Keep the Mosfet as high as possible above the PCB, with about 1mm of the leads protruding below the PCB.

Final assembly

The cutting and drilling guide (Fig.5) should help you to make the required cutouts in the case. You can download that as a PDF, along with the panel label artwork, from our website at https://pemag.au/Shop/11/6928

Fig.5 shows the locations, sizes and shapes of the IEC connector and GPO cutouts, which are suitable for all three enclosure options.

For the version that mounts on a wall for controlling ceiling fans, you don't need to make the IEC connector hole or the one for the lid-mounted GPO. Just fashion the cutouts for the switch, potentiometer and Earthing points.

As mentioned earlier, in the wall-mount application, the potentiometer can be mounted either on the PCB for end-mounted speed adjustment or on the lid. Regardless, the box must be Earthed. Access holes to fit grommets for the wiring can be made in the base of the box so that the fan wiring can be concealed in the wall.

For the standalone controller, first mark the hole position for the IEC

Parts List – Fan Speed Controller Mk2

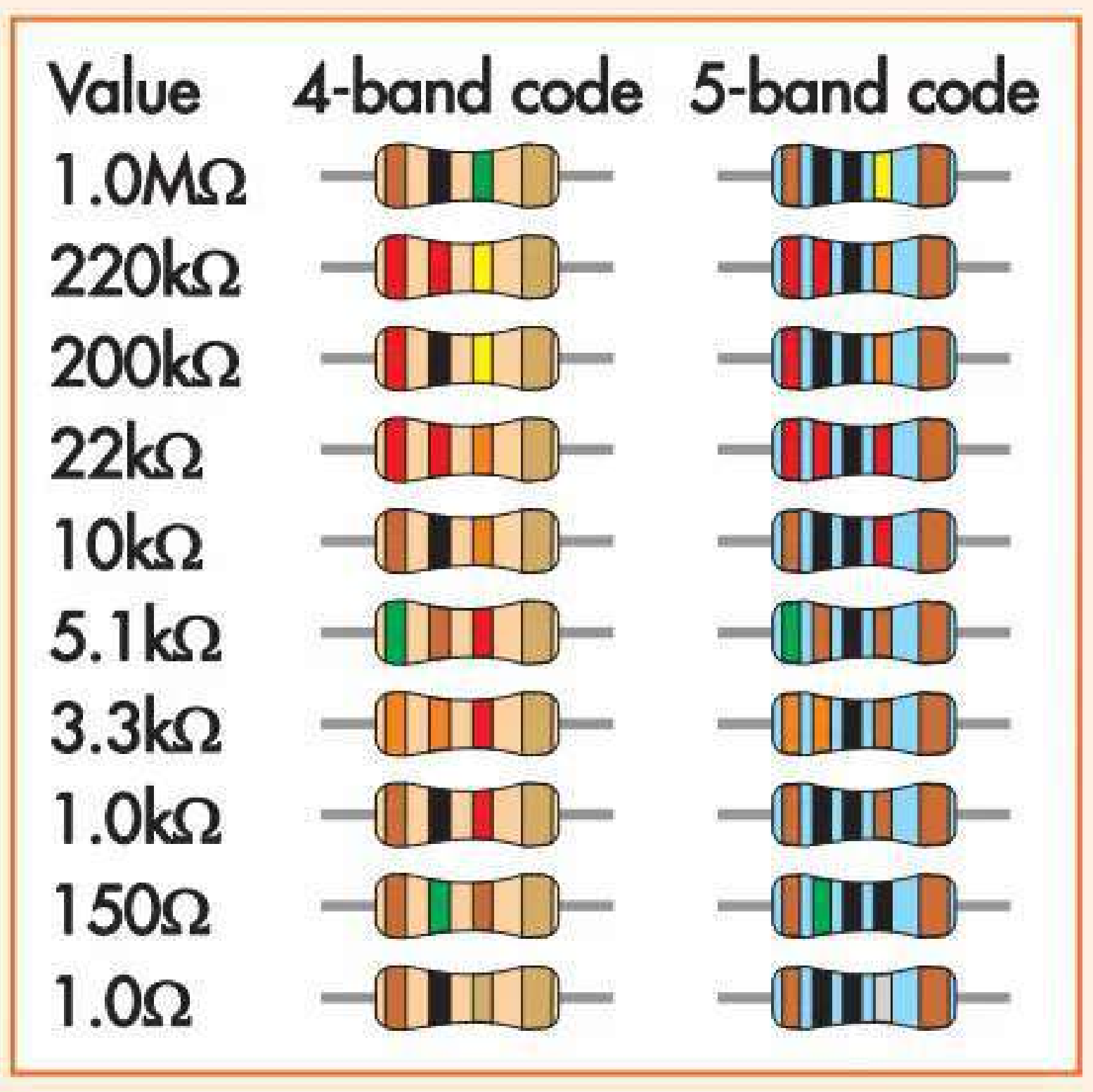
1 double-sided PCB coded 10104241, 94 × 79mm
1 115 × 90 × 55mm Gainta G113/BS25MF IP65/IP66 diecast aluminium box (recommended) **[TME G113 or BS-25MF]** ***OR***
1 119 × 94 × 57mm Bud Industries CU-234 economy diecast box **[Farnell, Mouser, DigiKey]**
1 panel label (see text)
1 10kΩ dual-gang 24mm PCB-mount linear potentiometer (VR1)
1 plastic knob to suit VR1
1 6-way 15A 300V terminal barrier strip, 8.25mm pin spacing (CON1)
1 SPST 10A 250V AC rocker switch with integrated neon lamp (S1)
1 normally-closed 10A 50°C thermal switch (TH1) **[Farnell 1006842]**
1 1A 250V AC M205 fuse (F1)
1 8-pin DIL IC socket (**optional**)

Semiconductors
1 LM358 dual single-supply op amp, DIP-8 (IC1)
1 400V 10A N-channel Mosfet, TO-220 (Q1) **[FQP11N40C (Farnell 2453436)]**
1 15V 1W zener diode (ZD1) **[1N4744]**
1 400V 6A PW04 diode bridge rectifier (BR1)
1 1N4004 1A 400V diode (D1)
1 1N4148 200mA 75V signal diode (D2)

Capacitors
1 470µF 25V 105°C PC electrolytic
2 10µF 16V 105°C PC electrolytic
1 10µF 50V 105°C non-polarised (NP) PC electrolytic
2 100nF 63V or 100V MKT polyester

Resistors *(all ¼W, 1% axial unless specified)*
2 1MΩ
1 220kΩ 1W 5%
1 200kΩ
1 22kΩ
1 22kΩ 5W
1 10kΩ
2 5.1kΩ
1 3.3kΩ
3 1kΩ
1 150Ω
1 1Ω 5W 5%

Hardware & cable
1 TO-220 mica insulating washer
1 TO-220 3mm screw hole insulating bush
4 5.3mm ID insulated quick connect crimp eyelets with 4-6mm wire diameter entry
1 200mm length of green/yellow striped 7.5A mains-rated wire
1 200mm length of brown 7.5A main-rated wire
1 200mm length of blue 7.5A mains-rated wire
1 160mm length of 5mm diameter heatshrink tubing
1 20mm length of 20mm diameter heatshrink tubing
2 M4 × 10mm panhead machine screws and hex nuts
2 4mm shakeproof (toothed) washers
3 M3 × 10mm panhead machine screws and hex nuts
3 extra 24mm potentiometer washers
10 100mm cable ties
2 M3.5 × 6mm screws (only for Jaycar HB5042 case)
4 M3 × 9mm Nylon spacers
4 M3 × 6mm panhead machine screws
4 M3 × 6mm countersunk head machine screws
1 small tube of thermal compound
1 2mm-thick piece of scrap aluminium sheet (if required; see text)



Value	4-band code	5-band code
1.0MΩ		
220kΩ		
200kΩ		
22kΩ		
10kΩ		
5.1kΩ		
3.3kΩ		
1.0kΩ		
150Ω		
1.0Ω		

Extra parts for the standalone version, for pedestal and box fans
1 13A chassis-mount mains socket [eg, RS 500-0459]
1 fused IEC mains input connector
1 7.5A IEC mains plug lead
2 M3 × 10mm countersunk machine screws and hex nuts
4 small stick-on rubber or felt feet

Extra parts for the wall-mounted version, for ceiling fans
1 M205 10A 250VAC panel-mount safety fuse holder
1-2 grommets or cable glands for input and output wires
1 600mm length of brown 7.5A mains wire (if VR1 is mounted on the lid)
1 120mm length of 5mm diameter heatshrink tubing (if VR1 is mounted on the lid)

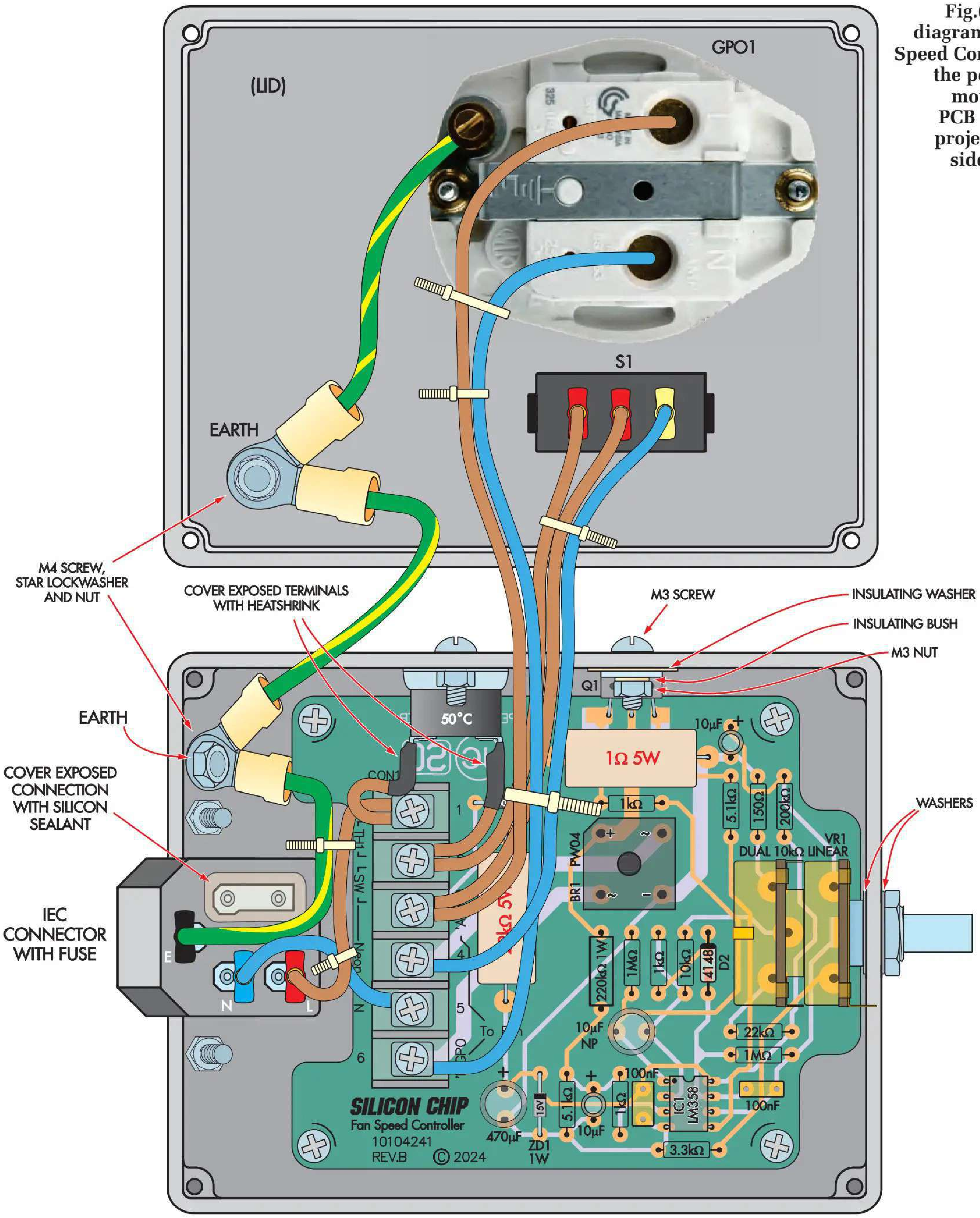

GPO1
(LID)
S1
EARTH
M4 SCREW, STAR LOCKWASHER AND NUT
COVER EXPOSED TERMINALS WITH HEATSHRINK
M3 SCREW
INSULATING WASHER
INSULATING BUSH
M3 NUT
EARTH
COVER EXPOSED CONNECTION WITH SILICON SEALANT
IEC CONNECTOR WITH FUSE
E
N
L
Q1
50°C
CON1
1
4
5
6
TH1
SW
N
To Fan
GPO
1Ω 5W
10µF
1kΩ
5.1kΩ
150Ω
200kΩ
WASHERS
VR1
DUAL 10kΩ LINEAR
BR1 PW04
220kΩ 1W
1MΩ
1kΩ
10kΩ
4148
D2
10µF NP
22kΩ
1MΩ
100nF
IC1 LM358
100nF
470µF
15V
ZD1 1W
5.1kΩ
10µF
1kΩ
3.3kΩ
SILICON CHIP
Fan Speed Controller
10104241
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Fig.6: the wiring diagram for the Fan Speed Controller with the potentiometer mounted on the PCB and its shaft projecting out the side of the case.

connector and Earth screw in the end wall of the case. The IEC connector mounts with a gap of about 4mm from the base of the case to the bottom of the IEC connector. The hole is made by drilling a series of small holes around the perimeter of the desired shape, knocking out the piece and filing it to shape.

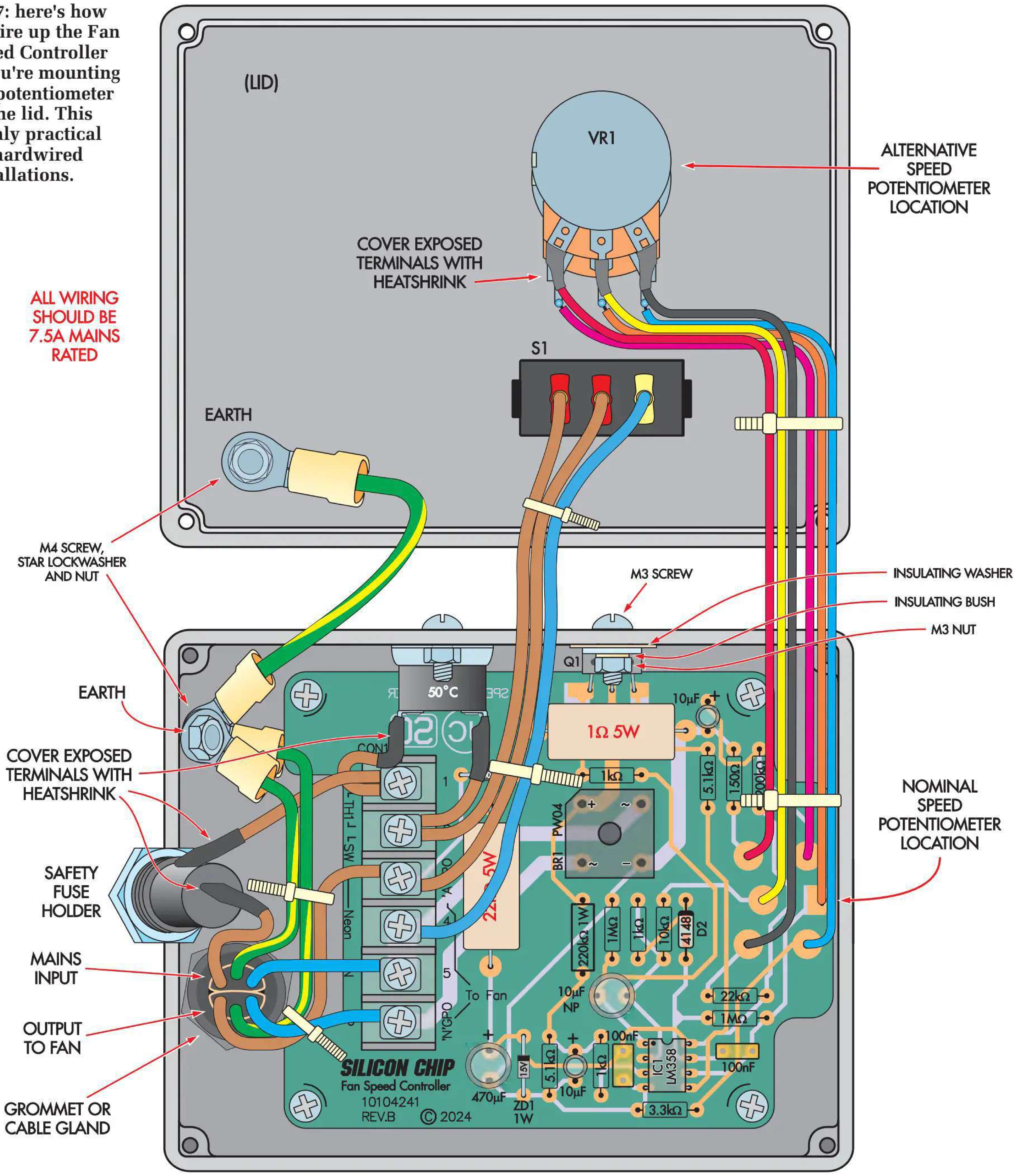
Alternatively, use a Speedbore drill to make a larger round hole to remove most of the required area, then file that hole to the required shape. The Earth screw hole is 4mm in diameter.

A hole is required for the potentiometer at the opposite end of the box. Measure the height of the potentiometer shaft above the base of the enclosure and mark out the drilling position at the end of the enclosure. Alternatively, for the potentiometer mounted on the lid, drill the hole in the centre of the GPO cutout.

Note that the potentiometer can only be installed on the lid for the ceiling fan version that doesn't require the GPO socket.

Insert the PCB into the case and note that the leads for Q1 must be kinked outward from the PCB a little so the

Fig.7: here's how to wire up the Fan Speed Controller if you're mounting the potentiometer on the lid. This is only practical for hardwired installations.


(LID)
VR1
ALTERNATIVE SPEED POTENTIOMETER LOCATION
COVER EXPOSED TERMINALS WITH HEATSHRINK
ALL WIRING SHOULD BE 7.5A MAINS RATED
S1
EARTH
M4 SCREW, STAR LOCKWASHER AND NUT
M3 SCREW
INSULATING WASHER
INSULATING BUSH
M3 NUT
Q1
50°C
EARTH
COVER EXPOSED TERMINALS WITH HEATSHRINK
SAFETY FUSE HOLDER
MAINS INPUT
OUTPUT TO FAN
GROMMET OR CABLE GLAND
NOMINAL SPEED POTENTIOMETER LOCATION
CON1
TH1
SW
Neon
N
'N'GPO
To Fan
1
4
5
1Ω 5W
22Ω 5W
10µF
1kΩ
5.1kΩ
150Ω
200kΩ
BR1
PW04
220kΩ 1W
1MΩ
1kΩ
10kΩ
4148
D2
10µF NP
22kΩ
1MΩ
100nF
100nF
IC1 LM358
5.1kΩ
1kΩ
10µF
3.3kΩ
15V
ZD1 1W
470µF
SILICON CHIP
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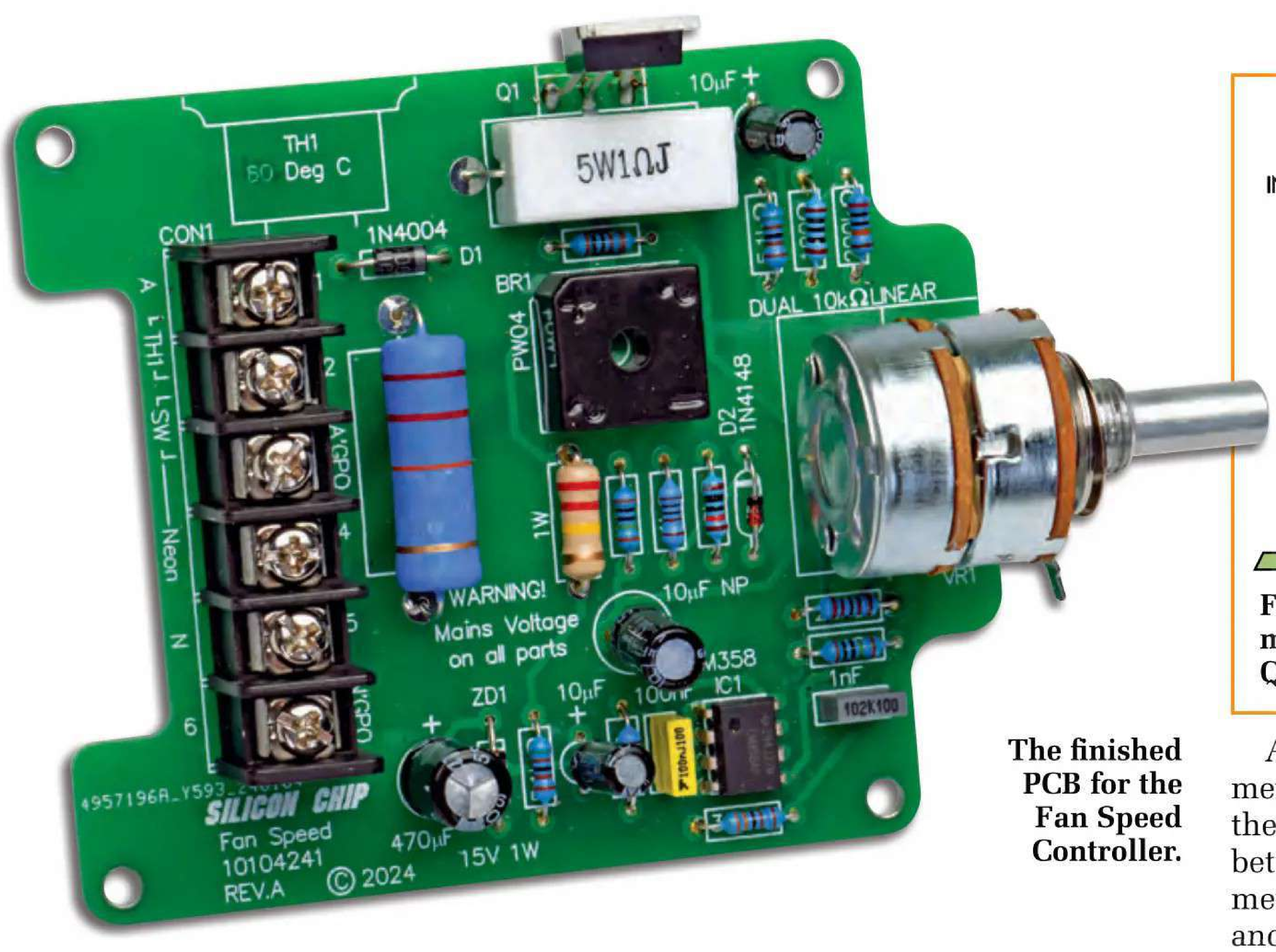
metal flange of the Mosfet sits in intimate contact with the side of the case.

You can then mark the mounting hole position for Q1's tab and drill it to 3mm in diameter. Deburr this hole on the inside of the case with a countersinking tool or larger drill to round off the sharp edge of the hole. This is to prevent punch-through of the insulating washer.

TH1 also mounts on the side of the box adjacent to Q1. There is room in the Jaycar HB5042 enclosure to mount TH1 against the side of the enclosure between two sets of protruding slots intended for mounting PCBs vertically. The Jaycar HB5064 enclosure does not have such slots, so there is plenty of room for mounting TH1.

For the Altronics case, there is insufficient room for TH1 to mount flat against the side of the enclosure. One solution is to grind away sufficient protruding slot material so the thermostat's body can sit flat.

The alternative is to make up an aluminium packing piece that's 19 × 45 × 2mm. This can sit between the protruding slots, and the thermostat can be mounted against that. In this



The finished PCB for the Fan Speed Controller.

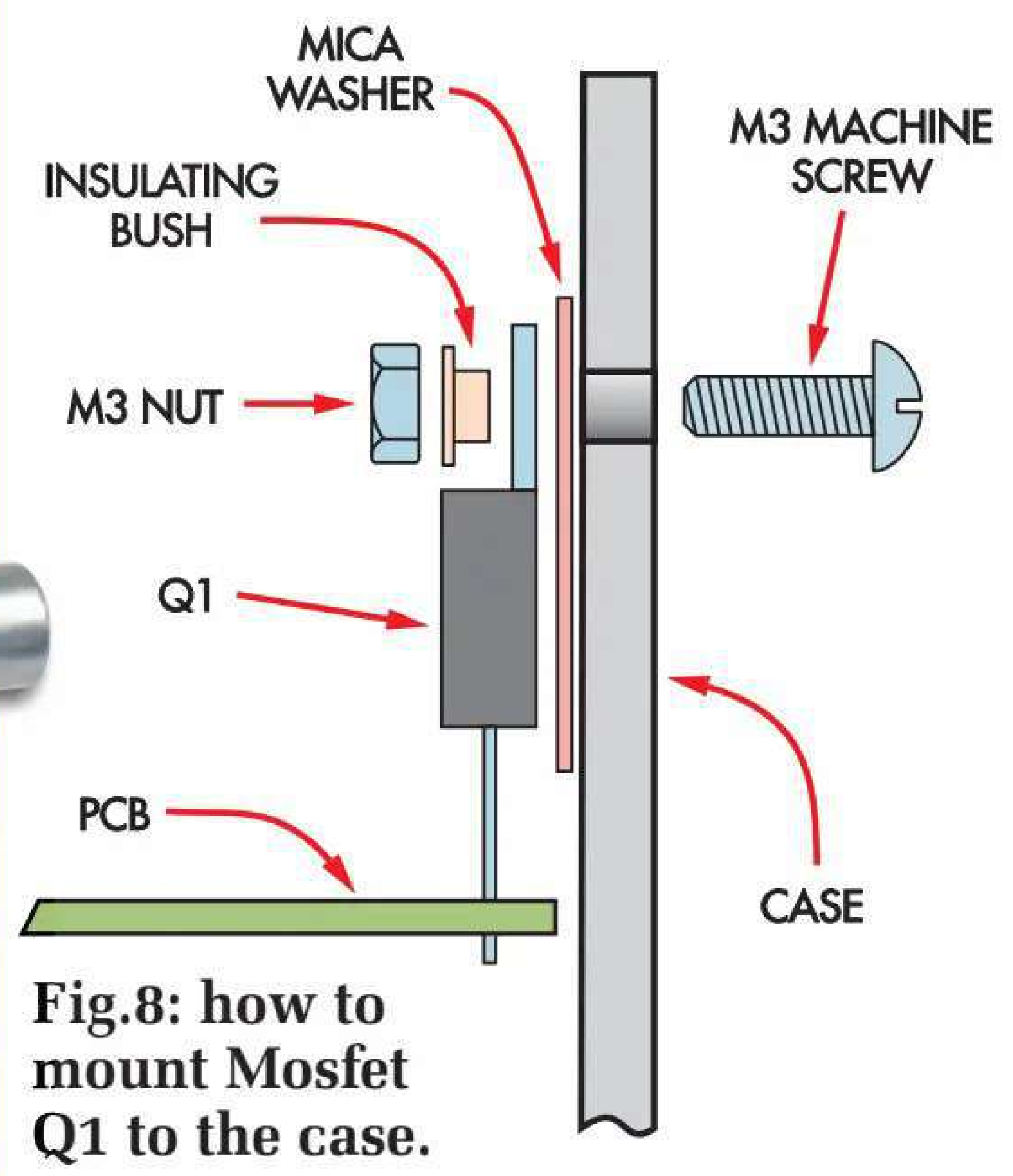


Fig.8: how to mount Mosfet Q1 to the case.

case, the top mounting hole should be about 8mm down from the top edge of the box.

Note that you will find it easier to install TH1 if the M3 nuts are tack-soldered to the thermostat mounting bracket. To do this, place the screws into the thermostat mounting bracket (when it is out of the case) and screw on the nuts, then solder them in place and remove the screws.

For the standalone version, holes are also required in the lid for the general purpose outlet (GPO) mains socket, the power switch and the Earth terminal. Four PCB mounting holes are also needed if you are not using the Jaycar HB5042 enclosure. The PCB is positioned so the speed potentiometer can protrude through the hole at the end of the enclosure.

Labels

Panel labels (see Fig.9) can be downloaded as a PDF from our website using the earlier link. Details on making a front panel label can be found at: https://pemag.au/Help/FrontPanels

The download includes two versions of the front panel. Which one you use depends on whether the control pot is mounted on the lid or is at the end of the enclosure.

If the potentiometer is PCB-mounted, its locating lug must be bent backward or snapped off, as we have not made a hole for it. Then slip three washers over the potentiometer shaft, insert it into the hole in the case by angling the board and drop the PCB onto the mounting points.

For the Jaycar HB5042 enclosure, secure the PCB to the case with the two screws supplied with the case plus two extra M3.5 × 6mm screws. For the other enclosures, the PCB is mounted using M3 × 6mm screws into M3-tapped standoffs. Secure the PCB-mounted potentiometer by placing another washer over the shaft on the outside of the case and doing up the nut on top.

Attach Q1 to the case with an M3 machine screw and nut, with the mica insulating washer and insulating bush as per Fig.8. Apply a thin smear of heatsink compound on all mating surfaces before assembly.

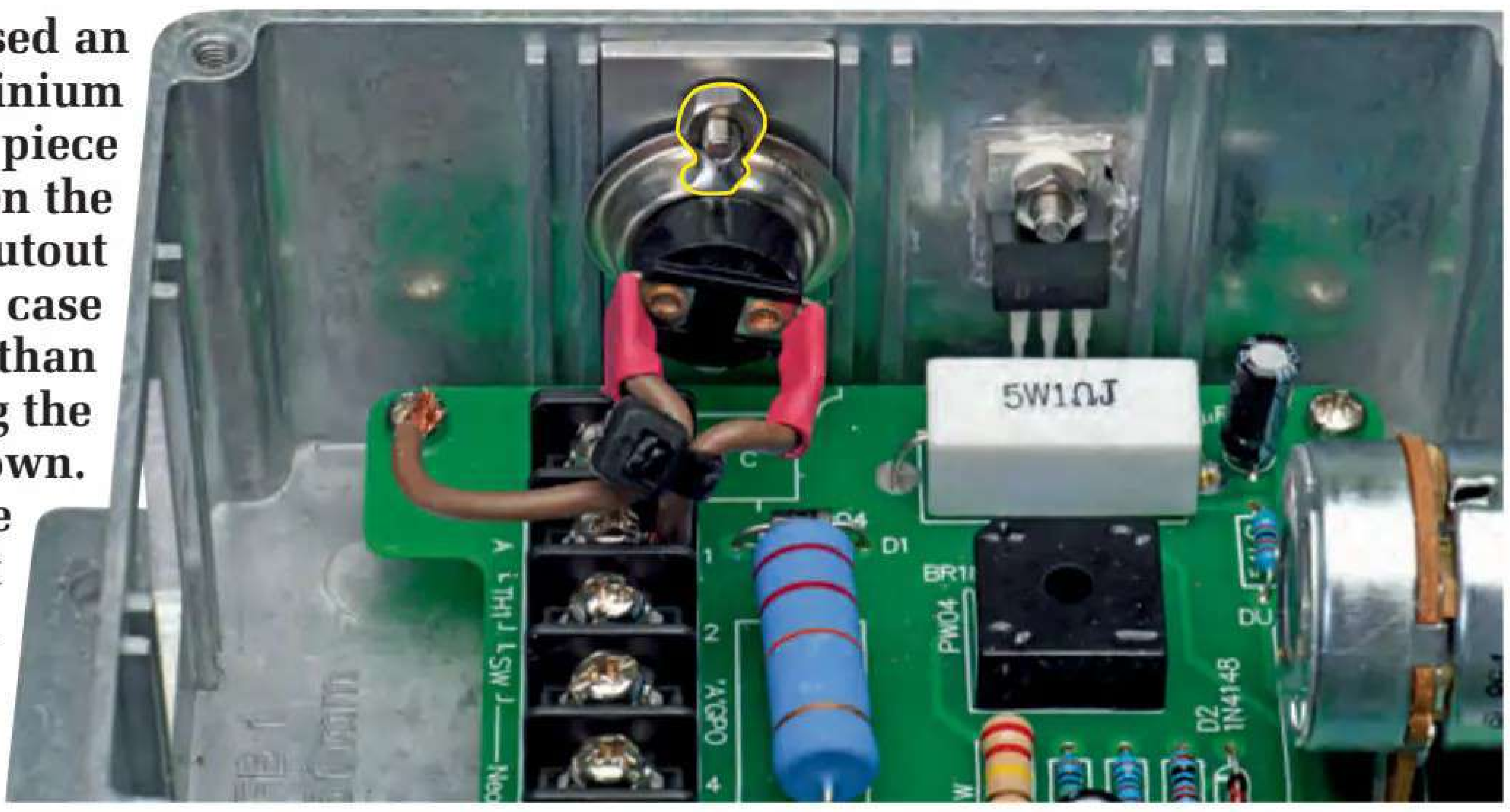
We use the mica washer in preference to silicone since mica has a higher thermal conductivity (lower °C per watt value), and the mounting screw can be tightened more. That keeps the Mosfet cooler compared to using a silicone washer.

After mounting Q1, check that the metal tab of the device is isolated from the case by measuring the resistance between them with a multimeter. The meter should show a very high resistance measurement (several megohms or possibly "0L") between the enclosure and Mosfet tab or the enclosure and any of Q1's leads. Check that it also reads close to 0Ω between the enclosure and the mounting screw.

The complete wiring diagrams for the two versions are shown in Figs.6 & 7. The Earthing details of the case are most important since Q1 and the potentiometer are all at mains potential, yet they are attached to the case. If the insulating washer or the insulation of the potentiometer were to break down, the case would be live (at 230V AC) if it was not properly Earthed.

The case lid must be independently Earthed rather than relying on the lid making contact with the base of the enclosure.

All mains wiring must be done using 7.5A minimum mains-rated (230V AC) wire. The IEC connector must be wired using the correct wire colours: brown for Live, blue for Neutral and



We used an aluminium packing piece between the thermal cutout and the case rather than grinding the rails down. Note the soldered nut highlighted in yellow.

green/yellow striped for the Earth. Live and Neutral wires soldered to the IEC connector must be insulated with heatshrink tubing covering all exposed metal.

Solder the Earth wire to the IEC connector Earth pin, ensuring the Earth terminal is heated sufficiently so that the solder wets and adheres properly to both the Earth terminal and wire. After that, use a crimping tool to secure the Earth wire into the crimp eyelet. Make sure the wire is securely crimped and can't be pulled out. The Earth wires from the Earth point to the lid and the GPO are also terminated with crimped eyelets.

Secure the Earthing eyelets with M4 machine screws, star washers and nuts. A second nut should be tightened on top of the first as a lock nut. The IEC connector is secured to the case by 10mm M3 countersunk head screws and nuts.

Finally, attach cable ties to hold the wire bundles together as shown in the wiring diagrams and the earlier photo of the fully assembled unit. Remember to place the four rubber feet on the bottom of the case.

Testing

As the whole circuit floats at mains potential, everything on the board should be considered unsafe to touch whenever the circuit is connected to the mains. That means the IEC mains power lead must be unplugged every time before opening the lid. Do not be tempted to operate the fan speed controller without the lid in place and screwed in position.

Before you power up the device, set VR1 fully anticlockwise. Also check all of your wiring very carefully against the overlay and wiring diagram. Verify that the case, lid and potentiometer are connected to the Earth pin of the power socket using a multimeter on its low ohms range.

If you are satisfied that all is correct, you are ready to screw the lid onto the case.

Note that the IP65 and IP66 enclosures are supplied with a rubber seal that goes between the enclosure base and lid. We did not use that seal so that heat from the case can transfer to the lid more efficiently for better dissipation.

The easiest way to test the circuit operation is to connect a fan. Apply power and check that you can vary its speed with VR1. Note that the fan controller box will begin to run quite warm with extended use when driving the fan at intermediate speeds. This temperature rise is normal. The temperature rise should be lower if the fan is set to a low speed.

Troubleshooting

If the speed controller does not work when you apply power, it's time to do some troubleshooting. First, a reminder: all of the circuitry is at 230V AC mains potential and can be lethal. That includes any exposed metal parts on components, except those tied to the Earthed case. Do not touch any part of the circuit when it is plugged into a mains outlet.

Before going any further, give your PCB another thorough check. Check for incorrectly placed components, incorrect component orientation or bad solder joints (dry joints, missed joints or bridges).

Optional heatsink

If the Fan Speed Controller works but cycles on and off due to the thermal cutout activating, a fan heatsink can be attached to the side of the enclosure where Q1 is mounted using M3 screws and nuts. A 105 × 25.5 × 55mm fan-type heatsink is recommended in this role.

The mounting holes are placed along the centre line of the heatsink. The lower hole should be positioned high enough not to foul the PCB when the nut is on.

The heatsink is positioned with its lower edge at the same level as the bottom edge of the box.

The heatsink should be counterbored at the Q1 and TH1 mounting screw positions. You can find where these screws are located by temporarily securing the heatsink onto the side of the case with the two M4 screws, with a thin layer of Blu-tack pressed onto the heatsink in each screw area. When the heatsink is removed, there will be an impression of the screw heads.

Drill out those two locations to a shallow depth using a larger drill to allow for the screw heads to sit inside the heatsink. Mount it with a smear of heatsink compound over the mating surfaces.

As an alternative, if countersunk screws are used for TH1 and Q1, there will be less counter-boring required on the heatsink.

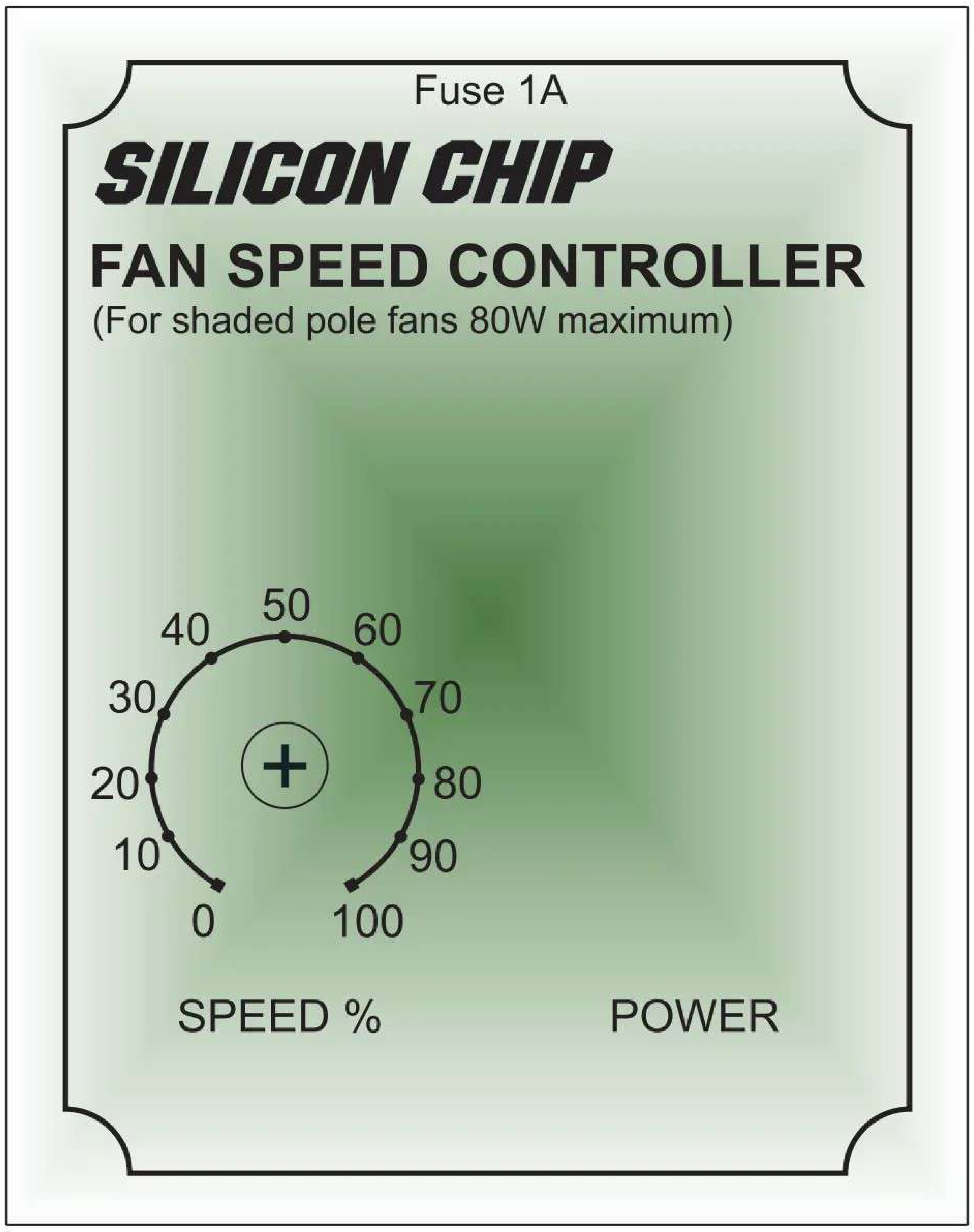
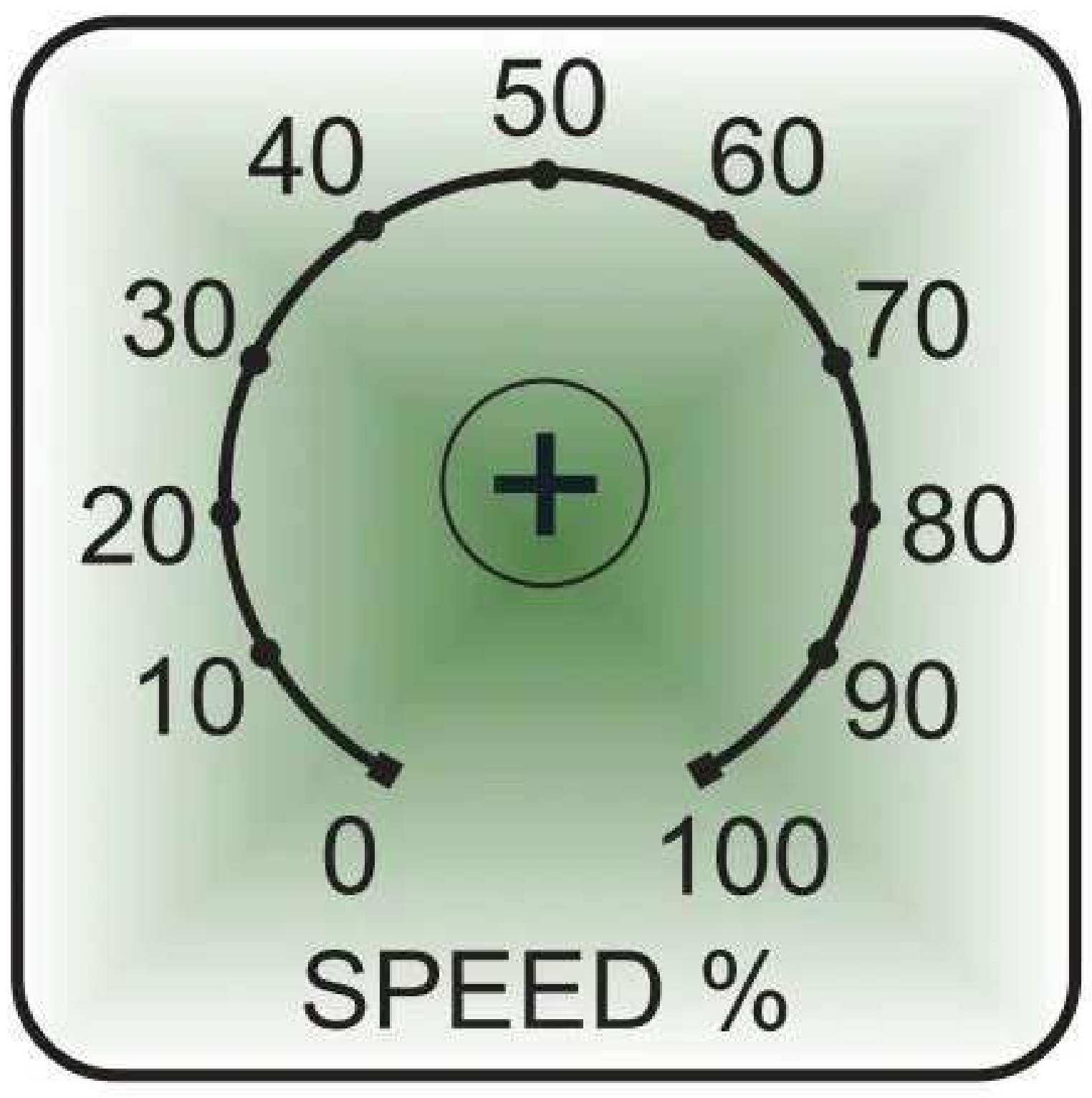
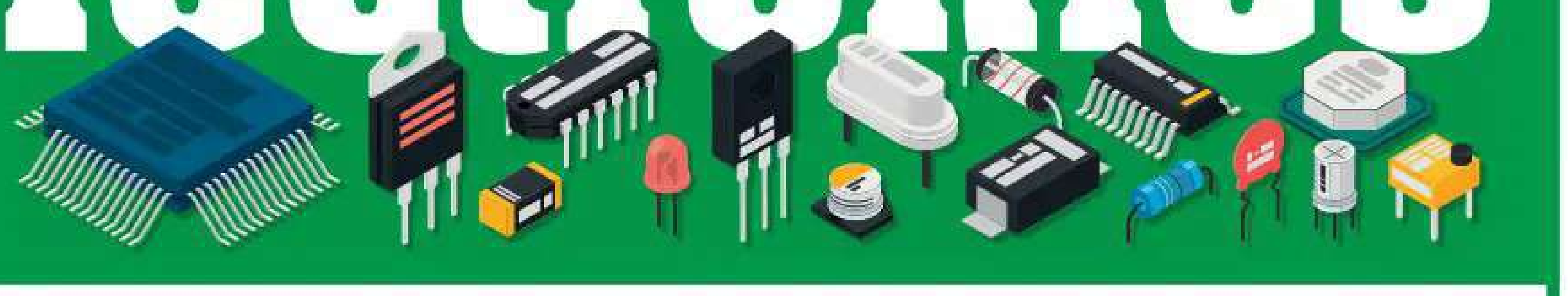


Fig.9: this label is for the Speed Controller with potentiometer on the lid. The other smaller label is only used if mounting the pot to the end of the case. All labels (including the alternative lid label) are available to download from https://pemag.au/Shop/11/6928



Precision Electronics

Part 4: Signal Switching



In this fourth article in this series, we will look at how to extend the current measurement range of the circuit we've been working on so far. To achieve that, we'll have to switch between two or more shunt resistors.

By Andrew Levido

In the previous article in this series, we developed our current-sense circuit (Fig.1) to the point where we could measure a 0–1A current in the high-side of a hypothetical power supply with a worst-case 25°C precision of around ±0.03%. Over the temperature range of 0–50°C, this error rose to just under ±0.2%.

That was the analog error only; it did not include any errors introduced by the analog-to-digital converter (ADC), which we will go into in a future article.

To achieve this level of precision, we were planning to apply a fixed gain calibration and a dynamic zero offset calibration in software, using the two switches shown in Fig.1. This level of precision would allow us to meaningfully measure current from 1A down to a few tens of milliamps, since our resolution is limited to ±2mA.

To achieve the microamp or better current sensitivity that we desire, we determined that we needed to switch in different shunt resistors to provide a series of current ranges.

So far, we have been using a 0.1Ω shunt resistor for the 1A range, which develops 100mV across it at full scale. This requires a differential-mode gain of about 25 to get our signal to a nominal 2.5V level for the ADC.

Assuming our power supply has some voltage headroom, there is nothing stopping us from increasing the shunt resistance by an order of magnitude, so it drops 1V at full scale. We can then decrease the gain to a nominal value of 2.5.

The power dissipation in the shunt resistor will increase accordingly, but any offset errors we see on the input side, including those that change with temperature, will be smaller in relation to the full-scale signal. That will improve the overall precision of the circuit. This will be important as the complexity – and therefore sources of uncertainty – of the circuit increases.

Table 1 (below) shows the ranges we could potentially implement, the current resolution we could expect, and the shunt resistors we would need for each one.

This table assumes we can maintain the ±0.2% error we have achieved so far. It suggests we should be able to realise our sub-microamp resolution ambitions if we can maintain a similar level of precision as we did with our previous efforts.

Before we get into the details of how we will switch the shunt resistors in and out, and the impact that will have on precision, we should look at the options available for signal switching. There are basically only two options:

1. We can use a mechanical switch such as a signal relay if we want to control it with a microcontroller.

2. Alternatively, we can use some form of electronic analog switch, which will most likely be based on field-effect transistors (FETs).

Signal relays

Signal relays are similar to power relays, but their design is optimised for low on-resistance and high linearity instead of power handling. They are usually rated for currents of 2A or less and for switching voltages under 50V. These aren't hard and fast definitions; there is plenty of grey area between the top end of signal relays and the bottom end of power relays.

Relays have the advantage of excellent on-resistance linearity with applied voltage and temperature. They have a very high off-resistance (essentially infinite) and virtually zero leakage since the switching path is electrically isolated.

Typical initial on-resistances for signal relays range from about 10mΩ to 200mΩ. The word "initial" is important here – the on-resistance of signal relays generally increases with the number of operations, as shown on the right side of Fig.2. This is an extract from a data sheet for Panasonic TQ-series relays, although all brands behave in more or less the same way.

It's also worth noting that the operating and release voltages, shown on the left, also worsen slightly with time

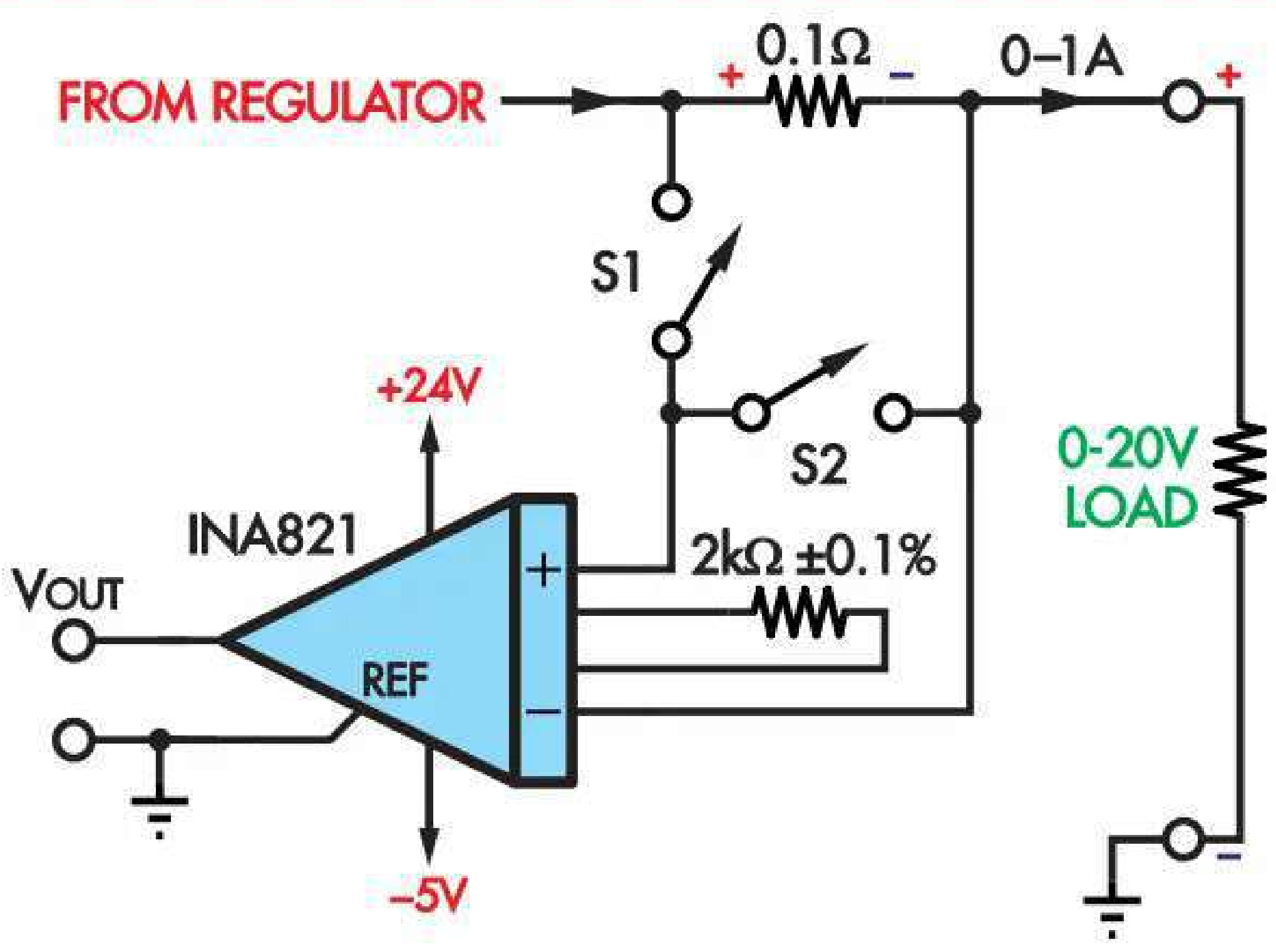


Fig.1: this is the circuit we designed last time. It is capable of a measurement resolution of a couple of milliamps; to measure lower currents, we need to switch ranges somehow.

Table 1 – current ranges using a fixed 0.2% error

Current range	Resolution (±0.2%)	RS (gain ≈ 2.5)
1.00A	±2.0mA	1.00Ω
100mA	±200μA	10.0Ω
10.0mA	±20μA	100Ω
1.00mA	±2.0μA	1.00kΩ
100μA	±200nA	10.0kΩ
10.0μA	±20nA	100kΩ
1.00μA	±2nA	1.00MΩ

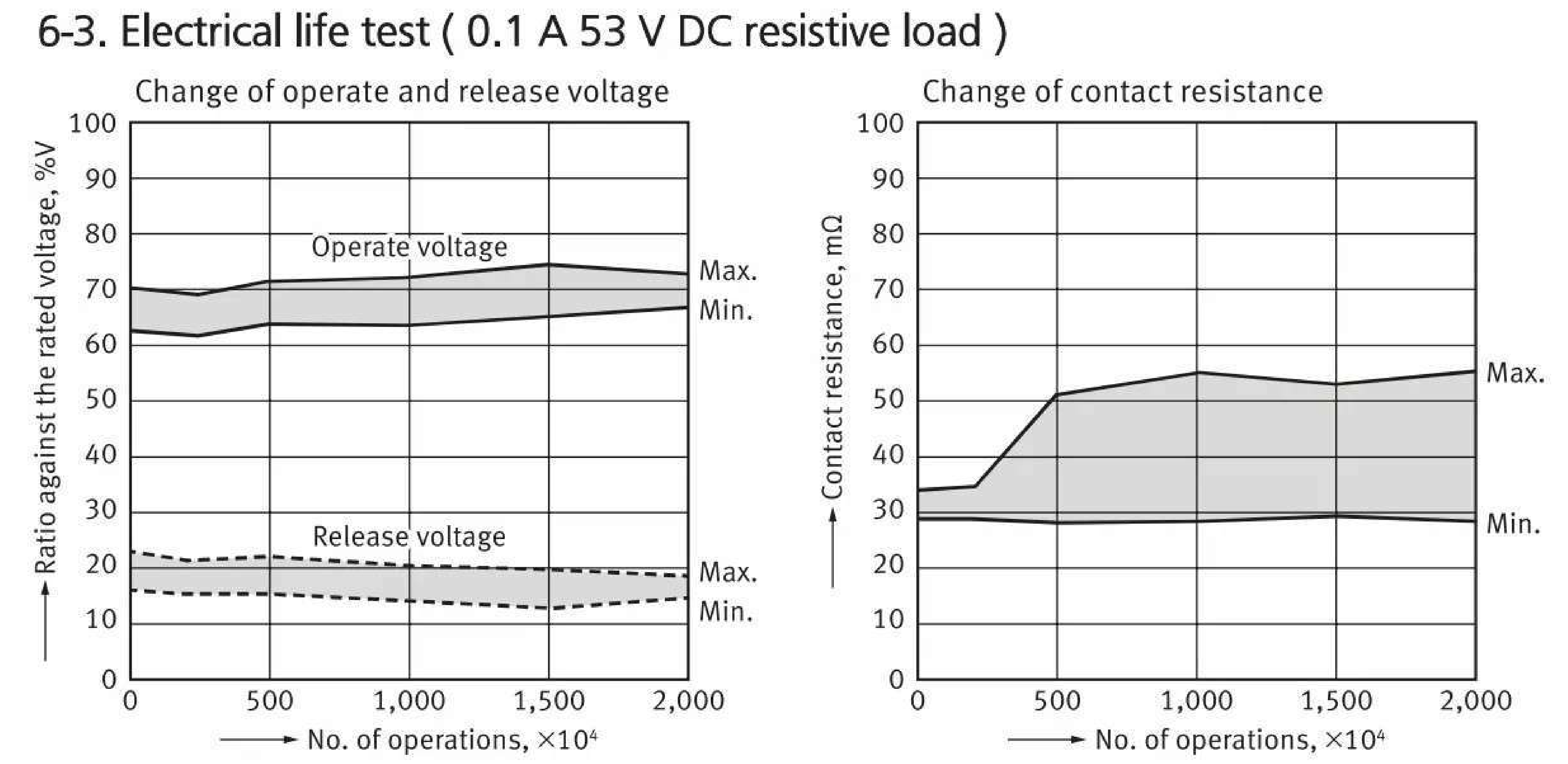


Fig.2: relays make great signal switches, but we should be aware that their contact resistance and operating voltages deteriorate with the number of operations.

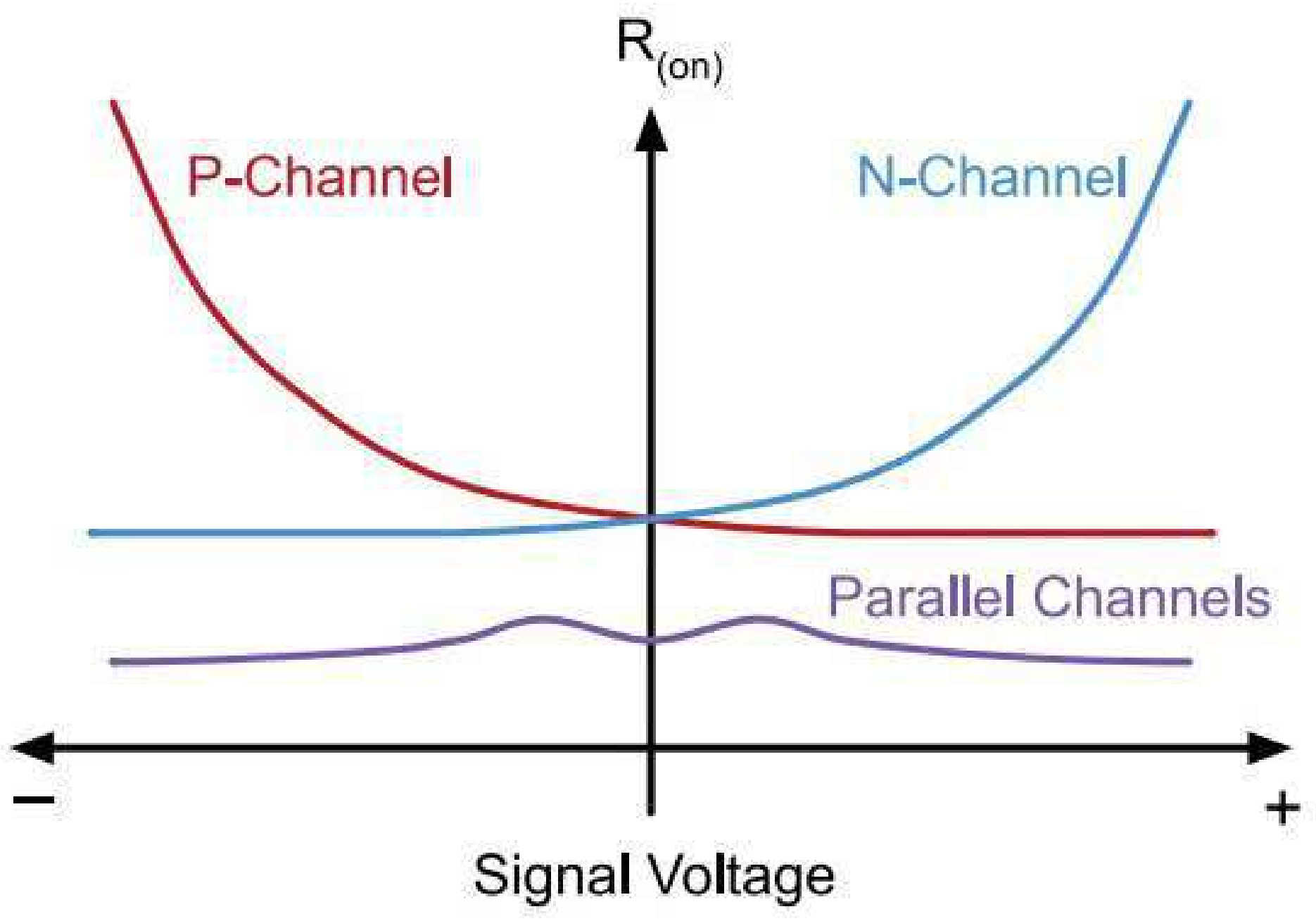


Fig.3: most analog switches use parallel N-channel and P-channel Mosfets to minimise the effect of the non-linear channel on-resistance of Mosfets.

as the relay's mechanical parts wear.

The failure rate data for the TQ relays suggests that 1% will have failed after 3.5 million operations and 10% after about 10 million operations when switching 5V at 1mA into a resistive load. That is a lot of operations, so it probably will not be of concern to the designer, but relays do have a limited life.

Panasonic deserves a lot of credit for publishing very comprehensive data for their relays. Not all manufacturers are this up-front in their data sheets.

Relays are not always good for very high-frequency applications, since their stray inductance and capacitance can be relatively high. Specialised high-frequency relays are available if you need them.

For precision circuits, we often use reed relays, which can have very low stray capacitance (0.5pF) and are available with internal electromagnetic screens which can help minimise induced noise or be used as a "guard" electrode when measuring minuscule currents. A reed relay is essentially a reed switch that's actuated by an electromagnet.

On the downside, relays are usually somewhat bulky and expensive, so designers tend to use them only when their unique characteristics are absolutely necessary. Instead, they generally use more compact and cheaper analog switches where they can (which offer the added benefit of an almost indefinite lifespan).

Analog switches

Analog switches are typically built from Mosfets since their drain-source resistance is controllable via gate voltage and the channel can conduct current in either direction. Because a Mosfet's channel resistance is non-linear with applied voltage, most analog switches use back-to-back N-channel and P-channel Mosfets.

The parallel on-resistance of the two devices is more linear than either one alone, as illustrated in Fig.3. The Mosfet substrates are connected to the analog power rails to maximise linearity.

By the way, if you are familiar with using discrete Mosfets as high-power switches, you may be puzzled by the comment that they can conduct current in either direction. That's because power Mosfets usually have an unavoidable 'body diode' in parallel with the channel in one direction, meaning they can only really switch current in one direction by themselves.

When fabricating multiple Mosfets on a single substrate as in a CMOS integrated circuit, the body diode is still there, but it is possible to choose where one end of that diode connects. Depending on what potential it is connected to, that body diode may never conduct under normal conditions, so it can effectively be ignored.

Thus, Mosfets in ICs (as well as the fairly unusual four-terminal discrete signal types that expose the bulk connection separately) can operate bidirectionally, similarly to JFETs.

The NMOS+PMOS architecture is used in switches such as those in the industry-standard DG41x series. Fig.4 shows the simplified circuit of one channel, extracted from the data sheet. As well as the back-to-back switching Mosfets, you can see a level shifter, which allows the control voltage (VIN) and logic supply (VL and GND) to be anywhere within the V+ to V– analog supply range.

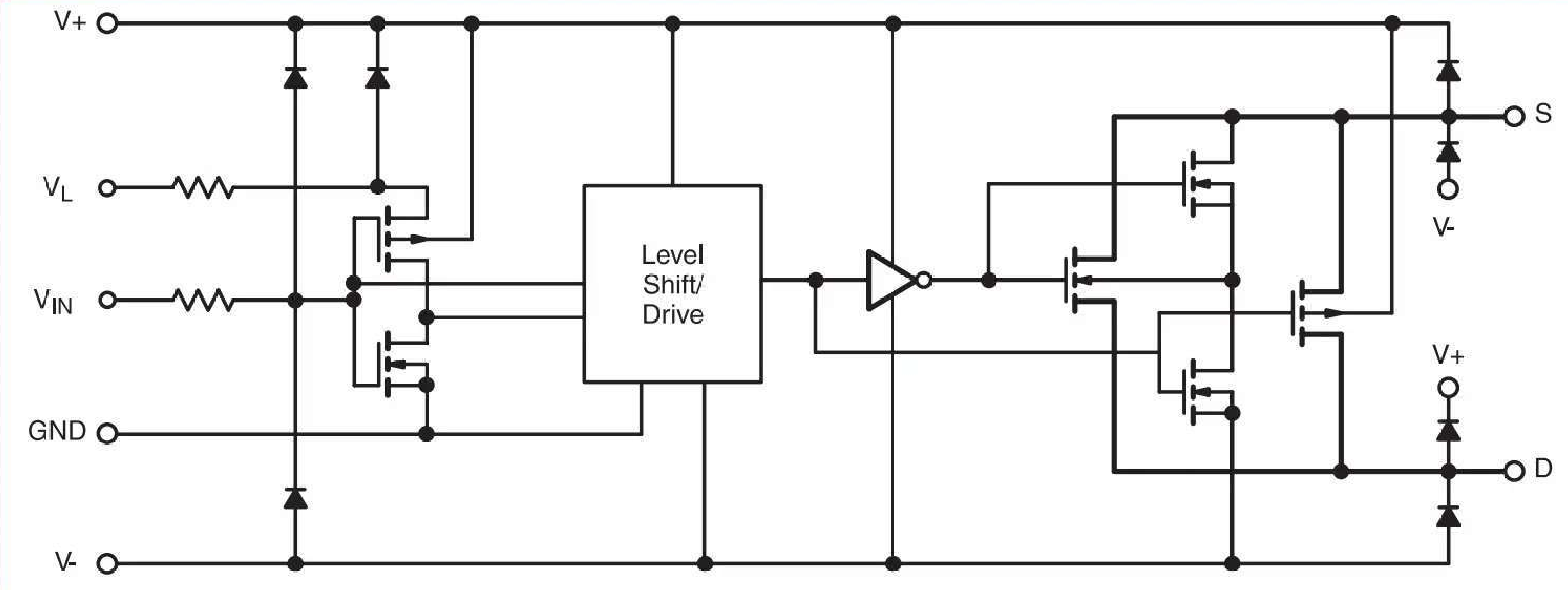


Fig.4: this simplified diagram of one switch from a DG41x series analog switch shows the parallel N-Channel and P-channel Mosfets. The level shifter allows the control signal and logic supply to be anywhere within the analog voltage range.

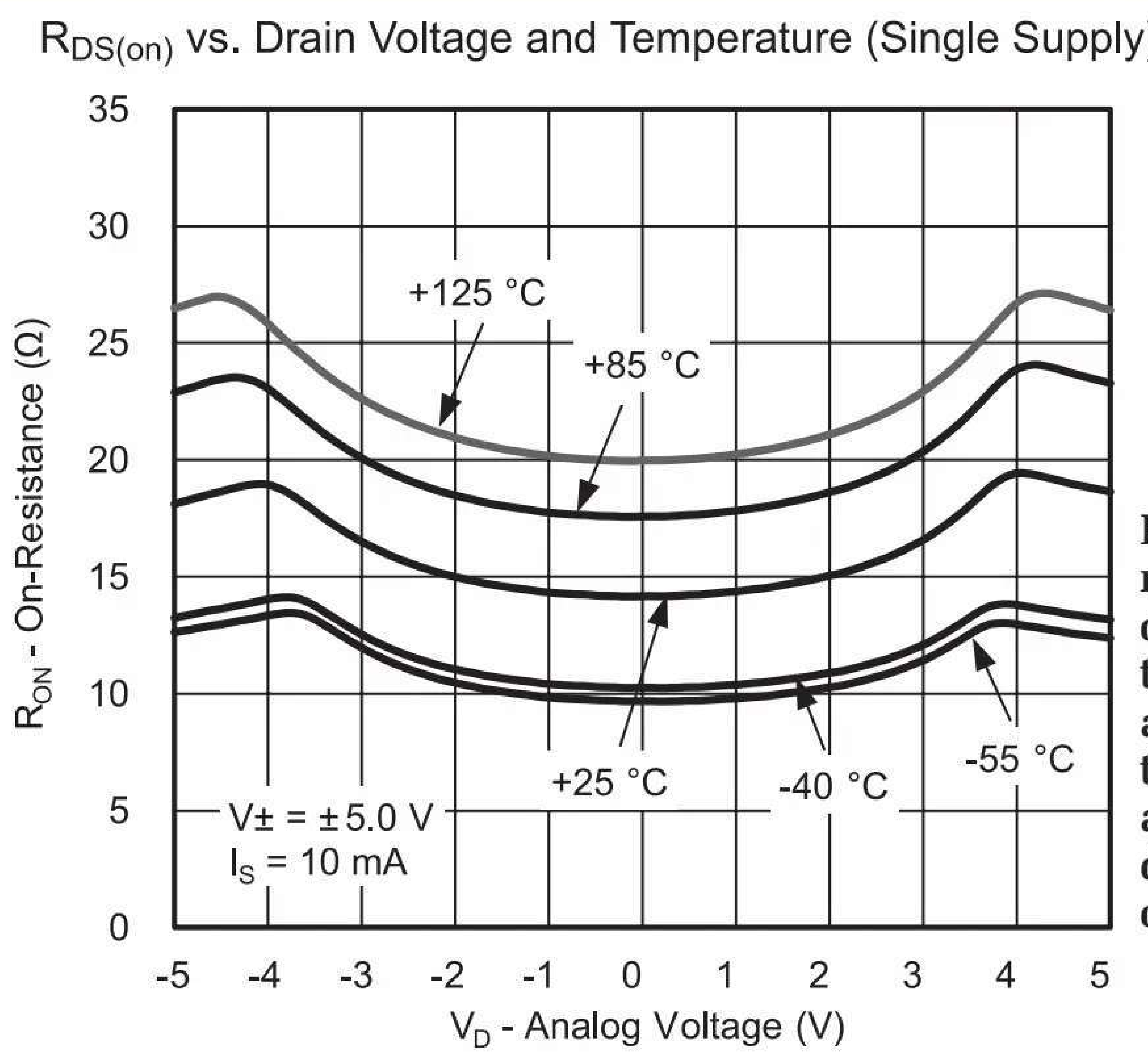


Fig.5: the on-resistance characteristic of this DG41x series analog switch shows the non-linearity and temperature dependence of the on-resistance.

The DG41x series switch on-resistance characteristic with ±5V rails is shown in Fig.5. The nominal on-resistance is anywhere between 10Ω and 20Ω, depending on temperature, and varies about 30% as the signal voltage changes.

The imprecision associated with analog switches can best be understood by looking at the on- and off-state equivalent circuits in Fig.6. In the on state (left), the on-resistance R_{ON} appears in series with the source resistance R_{SOURCE} to produce a voltage divider with the load resistance R_{LOAD}.

As we have seen, R_{ON} is non-linear and temperature-dependant, so the voltage error due to this divider will be uncertain. For this reason, we usually try to keep the load resistance as high as possible with respect to the sum of R_{SOURCE} + R_{ON}.

In the on state, a leakage current $I_{D(ON)}$ will produce a DC error voltage proportional to R_{LOAD} in parallel with R_{SOURCE} + R_{ON}. This can be minimised by keeping the source impedance as low as possible.

The channel capacitance $C_{D(ON)}$ will appear in parallel with C_{LOAD} and form an RC low-pass filter with R_{SOURCE} + R_{ON} – another reason to keep R_{SOURCE} low if you can.

In the case of the DG41x family of switches, R_{ON} can be up to 35Ω, $I_{D(ON)}$ can be up to ±15nA and $C_{D(ON)}$ is typically 35pF.

In the off state (shown in Fig.6), the leakage current $I_{S(OFF)}$ will produce a DC voltage across R_{SOURCE}, and $I_{D(OFF)}$ will produce a voltage across the load impedance, R_{LOAD}. The latter can be more difficult to manage, since we generally want to use a high load impedance for reasons described above. The DG41x switches have off-state leakage currents ($I_{D(OFF)}$ and $I_{S(OFF)}$) of up to ±15nA, and $C_{D(OFF)}$ can be up to 9pF.

Charge injection is another concern with analog switches, especially those with a low R_{ON} value. Achieving low R_{ON} requires physically large Mosfets, which have higher levels of gate capacitance. Whenever the gate of the Mosfet switches, this gate capacitance is charged or discharged via the drain and source. This means a charge is injected into the signal path when the devices switch.

The resulting voltage disturbance is a factor of the switch output and load capacitance, as shown in Fig.7. The charge is injected via C_Q and appears as a voltage spike or dip at the output, as $C_{D(ON)}$ in parallel with C_{LOAD} charge or discharge.

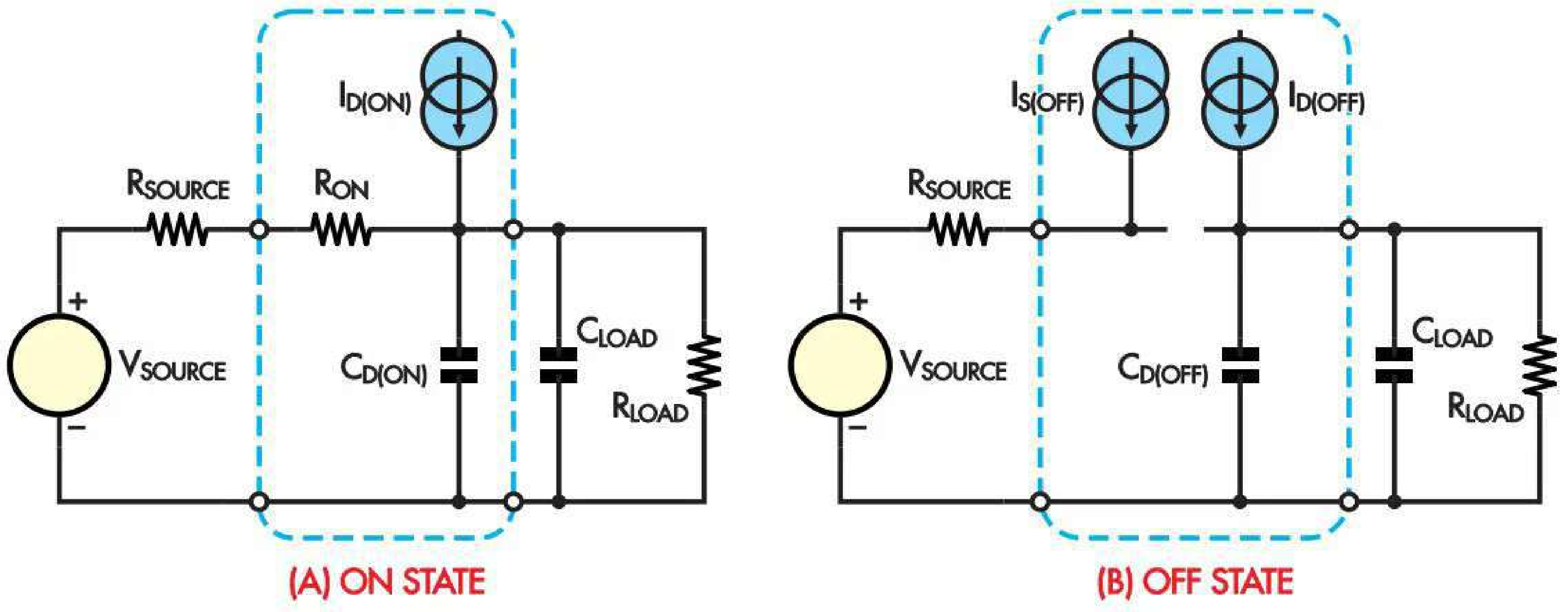
Each DG41x switch has a charge injection of 5pC. If the external load capacitance were 50pF, this would result in a voltage spike or dip of 59mV every time the switch changes state. This could very well create a significant 'pop' when switching audio signals – something to be aware of.

Of course, the input signal to this type of analog switch must stay within the power rails. For switches with back-to-back complimentary Mosfets, the signal voltage can extend all the way to both rails.

There are some newer analog switches with very good R_{ON} linearity. These appear to use a single N-Channel Mosfet with a very flat R_{ON} characteristic.

Fig.8 shows the on-resistance characteristic for one channel of the TMUX821x series of analog switches from Texas Instruments (TI). The on-resistance is very flat all the way from

Fig.6: these equivalent circuits show the leakage currents and internal capacitances present in analog switches in the on and off states.



(A) ON STATE

(B) OFF STATE

Fig.7: charge injection can cause voltage transients in the signal path when an analog switch is opened or closed. The effect is usually worse in low-$R_{DS(ON)}$ switches, where the gate capacitance (C_Q) is higher.

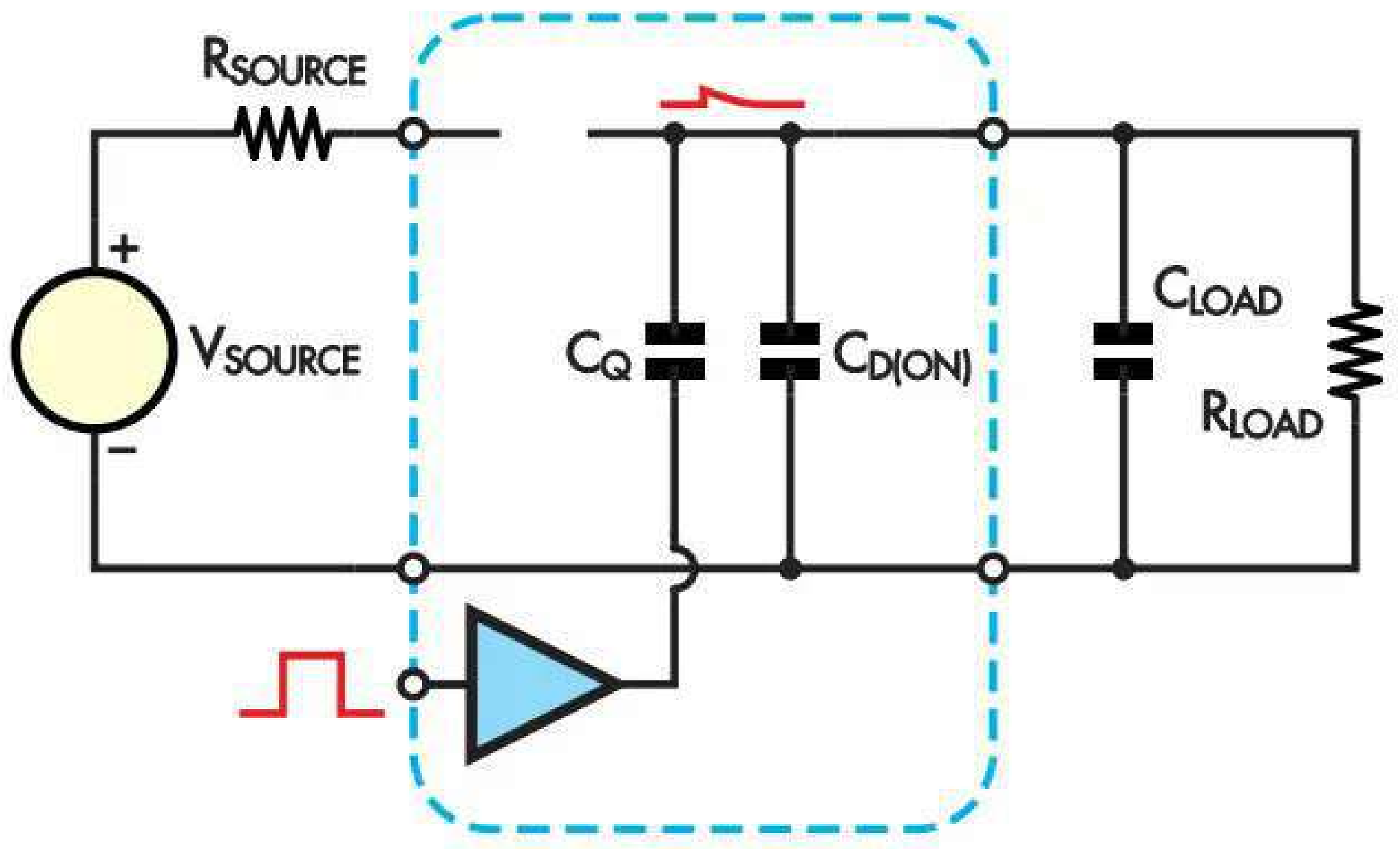
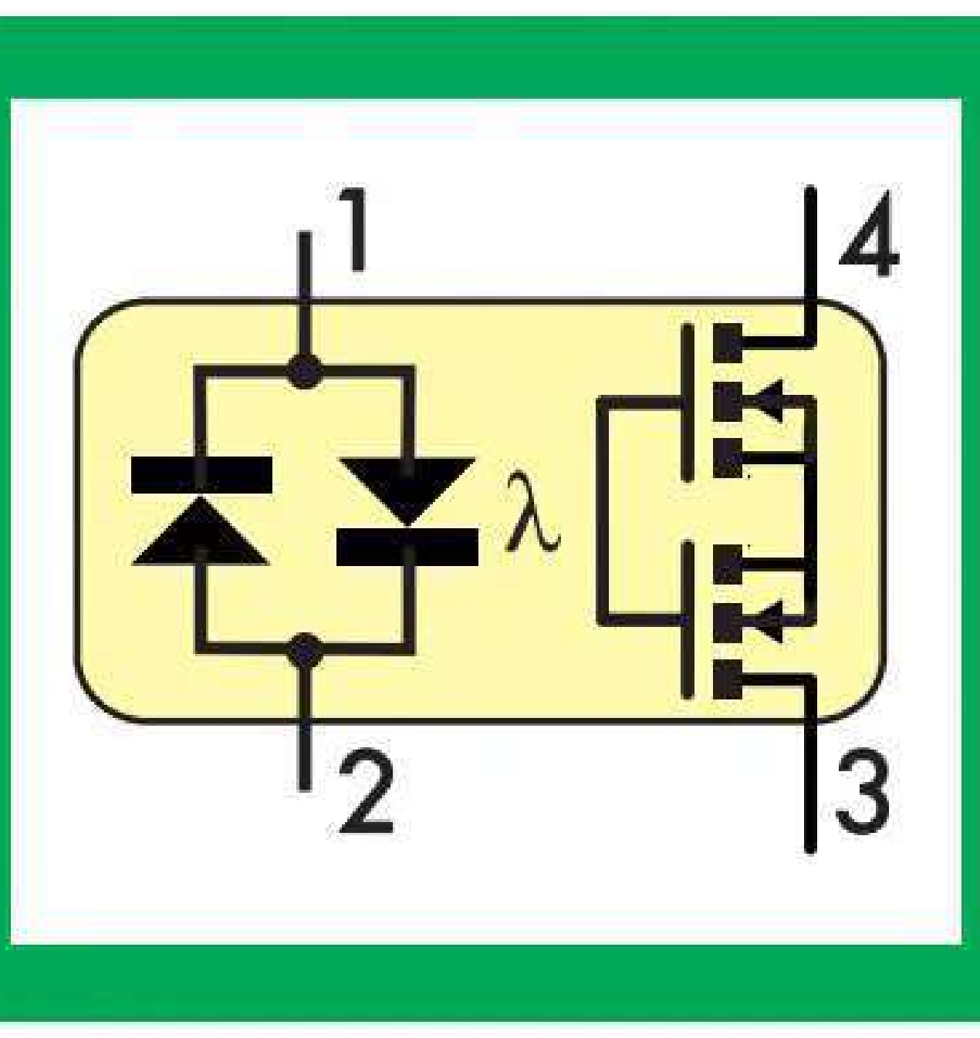


Fig.9: this interesting class of optically coupled analog switches may be suitable for some applications. They can switch a few hundred milliamps and provide good isolation between the control signal & switch.



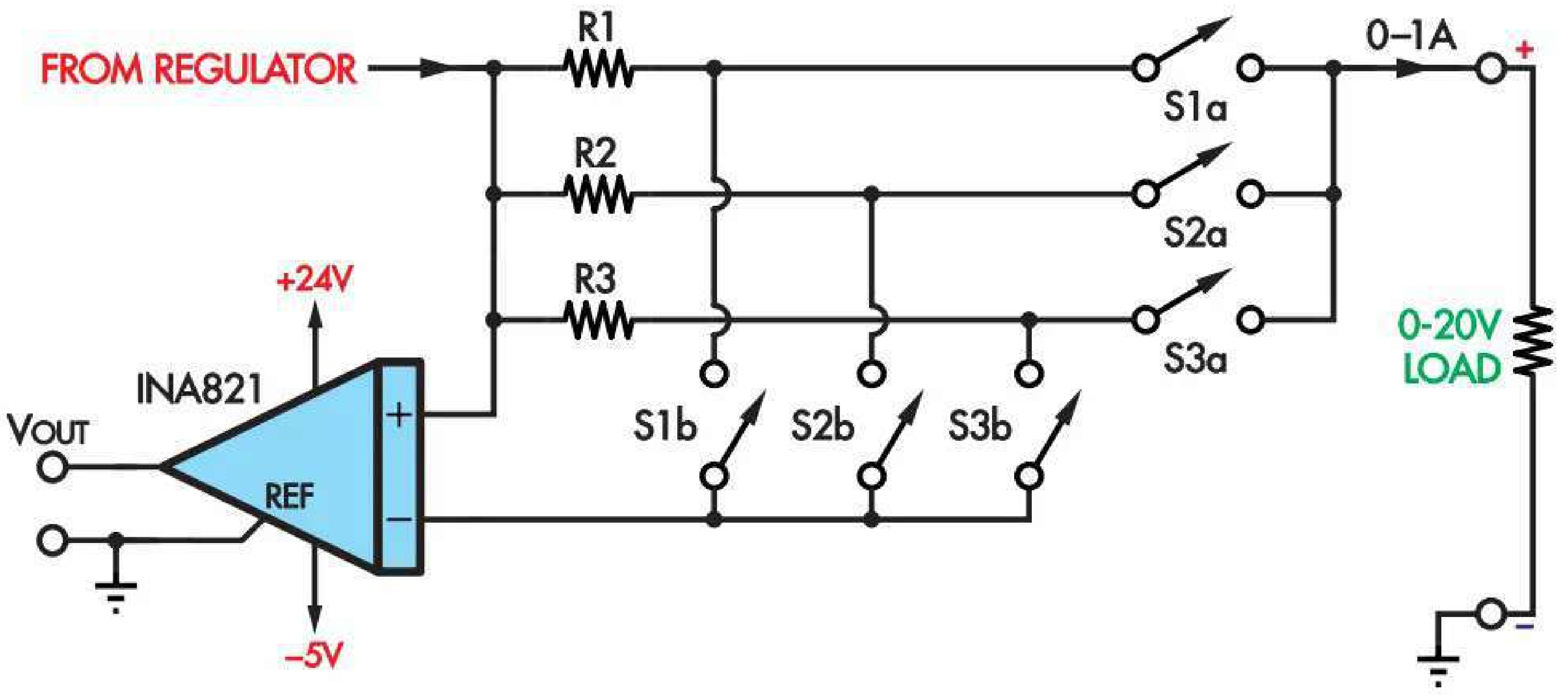


Fig.10: this circuit ensures the current-carrying switches (S1a, S2b and S3c) are not in the measurement path. That's helpful since the voltage drop across them is unpredictable. The shunt voltage sensing switches (S1b, S2b and S3b) carry no appreciable current, so the voltage drop across them will be minimal.

the negative supply up to a few volts short of the positive supply.

With the ±15V supplies shown here, the upper limit on signal voltage is around 10V to 12V, depending on how much non-linearity you can put up with.

Before we leave this discussion of analog switches altogether, I want to mention one more type that I have found useful in certain applications: optically coupled Mosfet switches, such as that shown in Fig.9.

These are a bit of a hybrid between relays, analog switches and optocouplers. They use inverse series Mosfets (for polarity independence), which are switched optically via an internal LED. A typical example, the AQY282GS, is rated for switching up to 60V (AC or DC) at 0.8A. It has a maximum on-resistance of 0.8Ω at 25°C, rising to twice that at 85°C.

The manufacturer does not provide any linearity data, but we can assume it will not be great.

They do have good input–output isolation (1000MΩ and 1.5pF), but up to 1μA of leakage between the output terminals when off. These devices are not super-fast – the switch-on time can be up to 5ms and switch-off up to 0.5ms. They are driven exactly like you would drive an optocoupler.

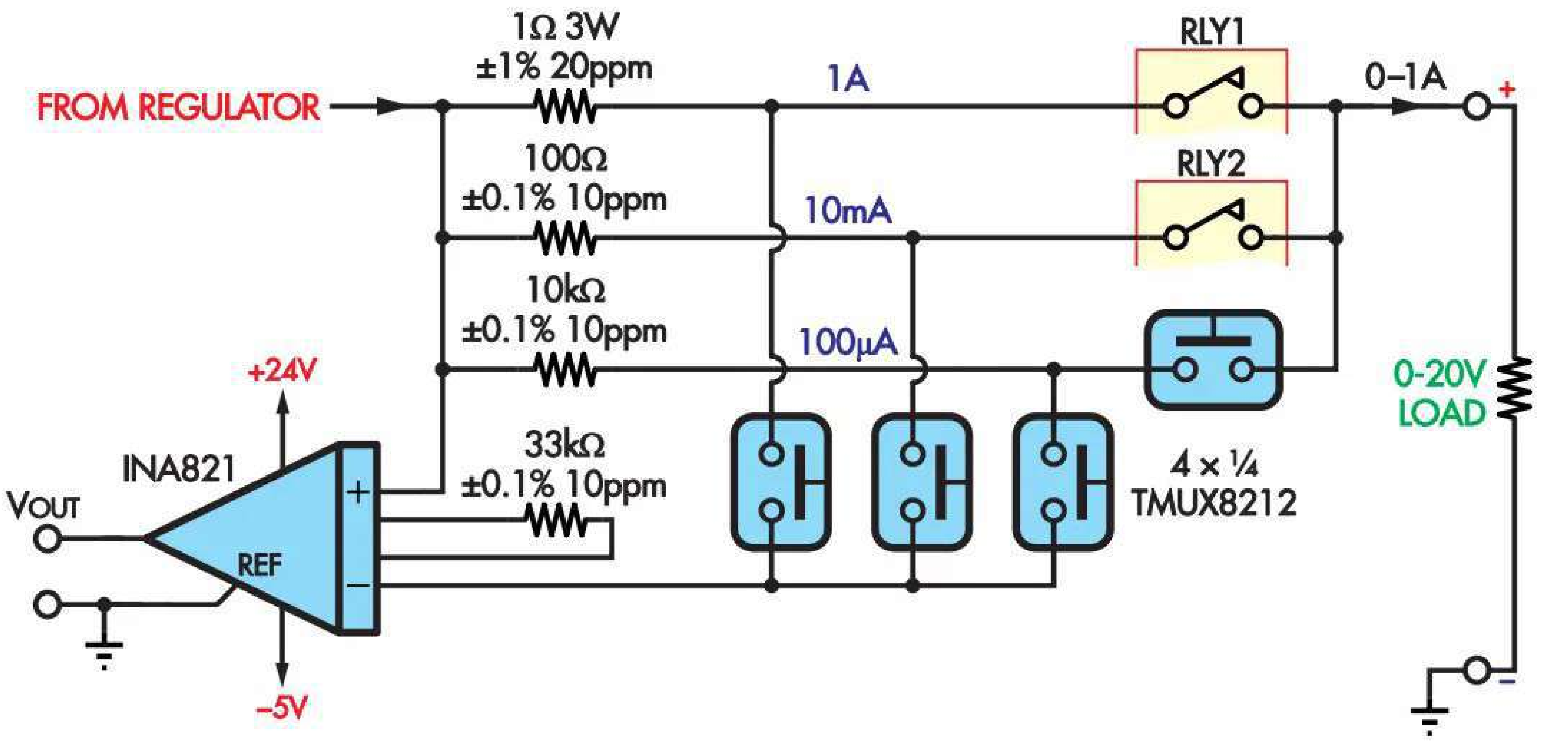


Fig.11: this circuit configuration was used to obtain the results described. Not shown are the DIP switches used to control the analog switches & relay coils.

Updating our design

So, armed with all this knowledge, how do we go about designing our multi-range current sensing circuit? Whatever type of switch we use to select the shunt resistors, it will add a material and unpredictable voltage drop.

We therefore can't just put the switching element in series with the shunt and measure the voltage across them both. Instead, we need to use the topology shown in Fig.10.

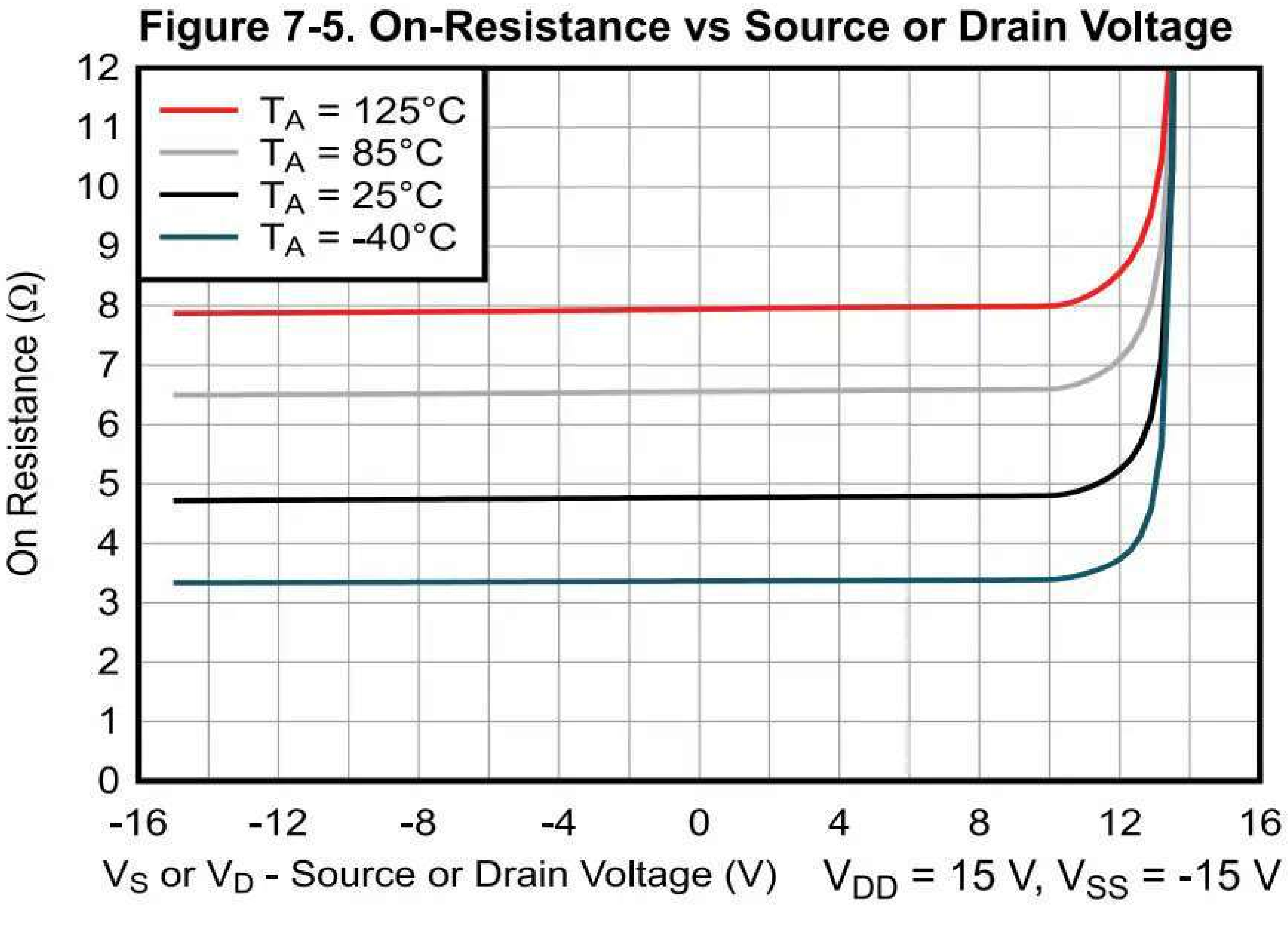


Fig.8: the on-resistance characteristic of the TMUX821x is remarkably flat for signal voltages from the negative rail up to a couple of volts short of the positive rail. This suggests a single Mosfet is being used. Note how the $R_{DS(ON)}$ is still highly temperature-dependent.

One of the "a" switches (S1a, S2a or S3a) is closed to select one of the shunt resistors, depending on the chosen current range. The corresponding "b" switch is also closed, connecting the relevant shunt resistor to the instrumentation amplifier's inverting input.

Since this input has a very high impedance, very little current flows through the "b" switch, so its on-resistance and non-linearity are largely irrelevant. The voltage drop across the active "a" switch, where appreciable current does flow, is not in the measurement path, so it does not impact the reading.

As a bonus, we get the zero-calibration state for free. If we close any "b" switch that does not have its corresponding "a" switch closed, we effectively short the inamp's inputs together via that shunt resistor, which will have close to zero voltage across it.

I decided to build a version of this circuit with 1A, 10mA and 100μA full-scale ranges. In a real application, you would probably implement

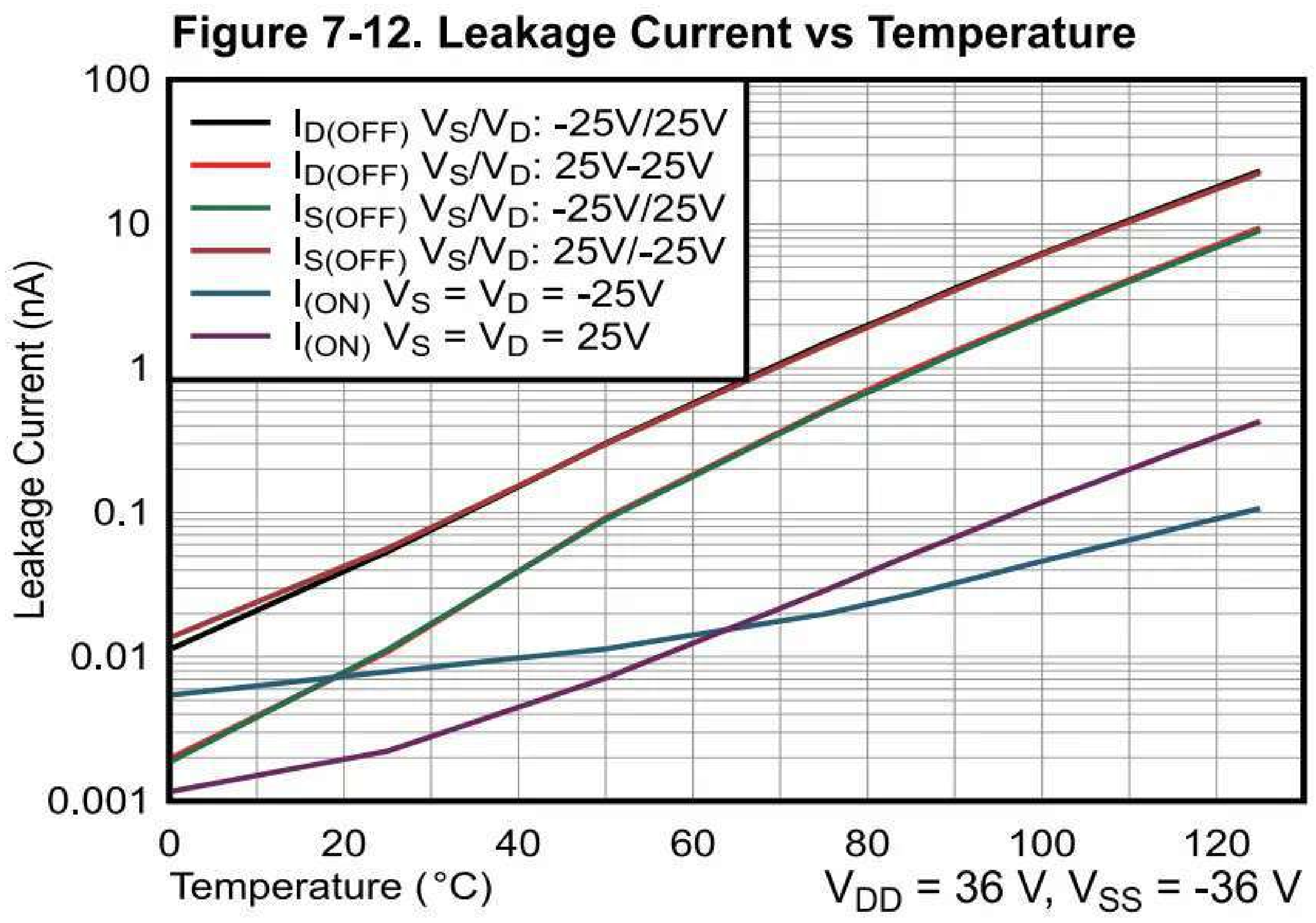


Fig.12: this graph, copied from the manufacturer's data sheet, shows the various leakage currents in the TMUX821x series of analog switches. As you would expect, they increase rapidly with temperature.

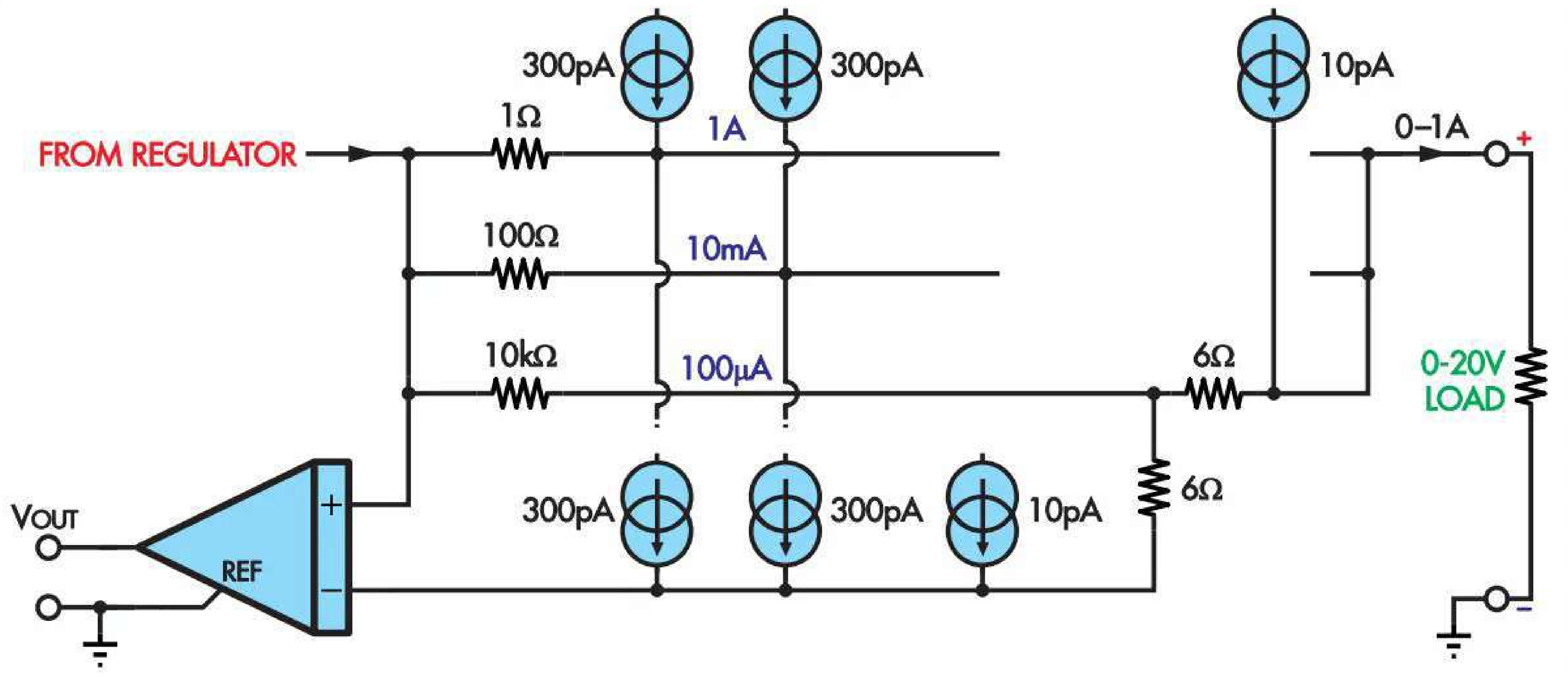
a range for each decade, but I wanted to keep things manageable for my experiments. I chose to use relays for S1a and S2a (the 1A and 10mA range respectively), although an analog switch could certainly be used for the latter range.

The 100µA range (S3a) and the three "b" switches used analog switches. This meant I could get away with just one quad analog switch package.

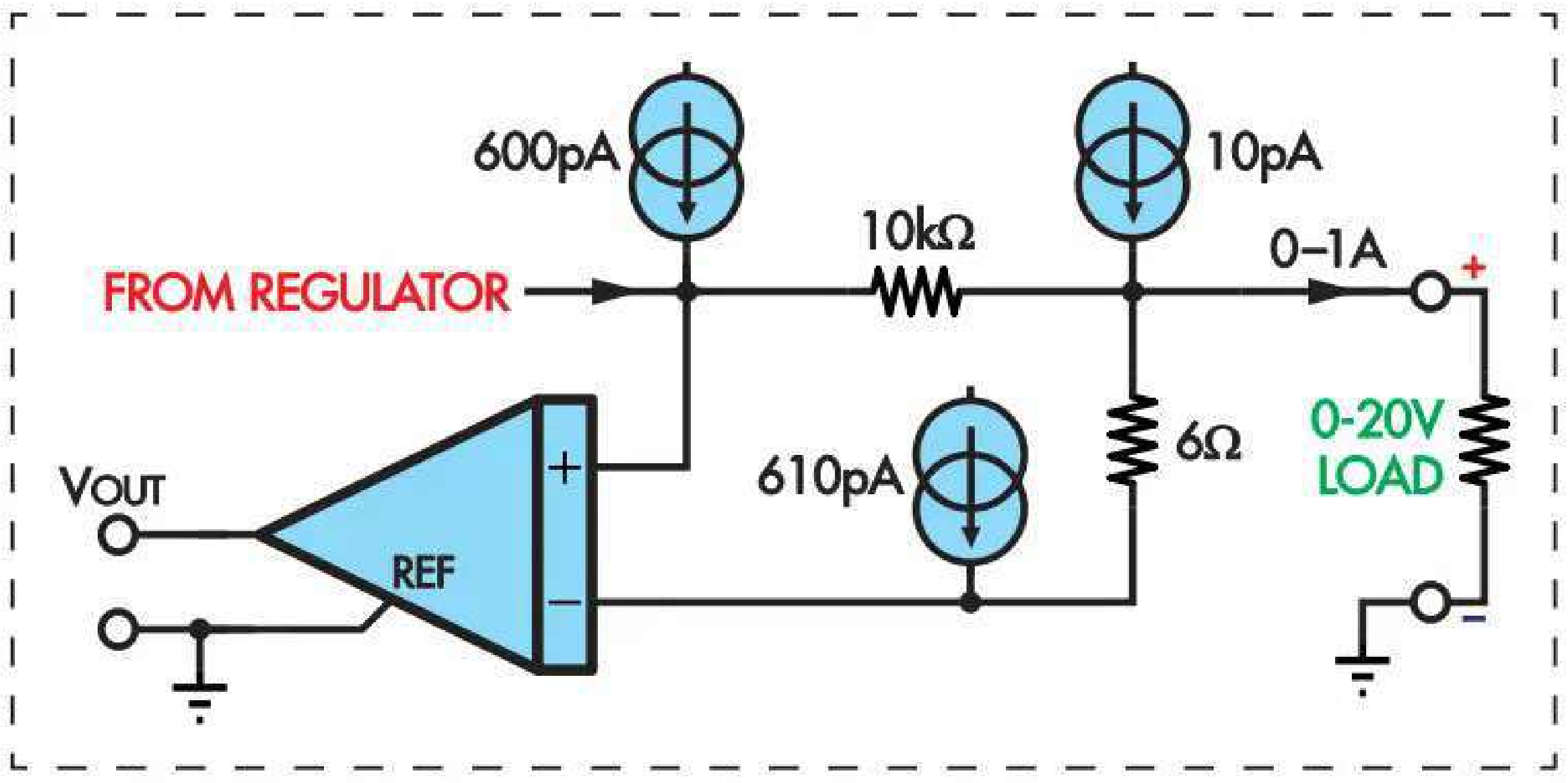
The key parts on the test board are shown in Fig.11. A 3.3V logic power supply and the dip switches driving the relays and analog switch control lines are not shown. I used a 1% tolerance 3W resistor for R1, since high-precision power resistors are super expensive.

I did, however, select a resistor with the best tempco (±20ppm/°C) that I could afford, since we can't trim out the temperature drift as easily as we can trim out the absolute resistance error. It is easier (and cheaper) to get high-precision 100Ω and 10kΩ resistors, so I chose devices with 0.1% tolerance and 10ppm/°C tempcos.

The relays I used were 3.3V coil 1A relays from Fujitsu's SY series that I happened to have on hand. The primary concern with selecting the analog switch was to get a unit with a sufficient voltage rating, since the supply voltages would be +24V and -5V, giving a total supply span of 29V. DG41x-series switches are limited to a supply voltage span of 12V.



Figs.13 & 14: the voltage error due to analog switch leakage is calculated by substituting the on and off equivalent circuits. As discussed in the text, the 600pA source can be ignored but the other two will cause an error. This diagram shows the 100µA range where the error is worse than the others. The simplified version is shown at right; it summarises the sources of error.



The TMUX821x range is good to ±50V, which is more than enough. The TMUX8212 includes four independent normally open switches, which is perfect.

From Fig.8, we can see that the analog switch on-resistance is under 5Ω at room temperature, with about ±1Ω change over the 0°C to 50°C range we are designing for. Fig.12 shows the leakage currents. At 50°C, the worst case for our design, $I_{D(ON)}$ is ±10pA or less, while $I_{D(OFF)}$ and $I_{S(OFF)}$ are each less than ±300pA.

Those figures are for ±36V supplies, so with our lower supply voltages, the values we experience are likely to be lower. However, in the absence of more detailed data, we have little choice other than to use those figures.

I used the cheaper of the two instrumentation amplifiers that we tested last time, the INA821, but this time with the gain set to about 2.5. Like last time, the op amp is powered from +24V and -5V rails.

Error budget

The easiest way to manage the error budget for a circuit with several configurations like this one is to calculate a separate budget for each range. The process is exactly the same as for the examples we created in previous instalments, except for the errors introduced by the analog switches.

We can distil the impact of the analog switches down to a single voltage error by substituting them with their equivalent circuits, as shown at the top of Fig.13.

Here, the circuit is shown with the 100µA range active (with the two analog switches closed and both relays open).

Fig.14 shows the same configuration with the leakage current sources consolidated. The 1Ω and 100Ω resistors disappear, since they are in series with current sources, which themselves have very high (theoretically infinite) source resistances. This simplification leaves us with three potential sources of leakage-induced voltage error.

The 600pA current feeding into the power rail on the source side of the shunt resistor can be ignored, since this current must flow either back into the regulator (where it does not matter), or through the shunt to the load (where it will be measured as part of the load current).

The 10pA source on the load side of the shunt will cause an error since this current can flow into the load

Table 2: 100µA range

		At Nominal 25°C		Additional error over 0-50°C (Nominal ±25°C)	
Error	**Nominal Value**	**Abs. Error**	**Rel. Error**	**Abs. Error**	**Rel. Error**
Shunt Resistor: ERA-6ARB103V (±0.1%, 100ppm/˚C)	10kΩ		0.10%		0.025%
Input voltage error due to shunt	**1V**	**1mV**	**0.10%**	**0.25mV**	**0.025%**
Input voltage error due to switch leakage	0V	6.2µV	0%	0mV	0%
Input voltage error due to bias (I_{OS} ±0.5nA, ±20pA/˚C)	0V	5µV	0%	5µV	0%
InAmp: INA821 (V_{OS} ±35µv, 5µV/˚C)	0V	35µV		125µV	
InAmp Input Voltage error total (Sum of Lines 2-5)	**0V**	**1mV**	**0.10%**	**0.380mV**	**0.038%**
InAmp Gain Resistor R_G: ERA-6ARB333V (±0.1%, 10ppm/˚C)	33kΩ	33Ω	0.10%	8.3Ω	0.025%
InAmp Gain Error (0.015% ±35ppm/˚C)			0.02%		0.088%
InAmp Gain (Line 7 × Line 8)	**2.5**	**0.0029**	**0.12%**	**0.0028**	**0.113%**
V_{OUT} DM (Line 6 × Line 9)	**0V**	**5.5mV**	**0.22%**	**3.8mV**	**0.151%**
V_{OUT} CM (20V, 100db, ±1.5db over 0-50˚C)	**0V**	**200µV**		**37.7µV**	
V_{OUT} (Line 10 + Line 11)	**0V**	**5.7mV**	**0.23%**	**3.8mV**	**0.152%**

Table 3: 10mA range

		At Nominal 25°C		Additional error over 0-50°C (Nominal ±25°C)	
Error	**Nominal Value**	**Abs. Error**	**Rel. Error**	**Abs. Error**	**Rel. Error**
Shunt Resistor: ERA-6ARB101V (±0.1%, 10ppm/˚C)	100Ω		0.10%		0.025%
Input voltage error due to shunt	**1V**	**1mV**	**0.10%**	**0.25mV**	**0.025%**
Input voltage error due to switch leakage	**0V**	**95nV**	**0%**	**0nV**	**0%**
Input voltage error due to bias (I_{OS} ±0.5nA, ±20pA/˚C)	**0V**	**50nV**	**0%**	**50nV**	**0%**
InAmp: INA821 (V_{OS} ±35µv, 5µV/˚C)	0V	35µV		125µV	
InAmp Input Voltage error total (Sum of Lines 2-5)	**0V**	**1mV**	**0.10%**	**0.3751mV**	
InAmp Gain Resistor R_G: ERA-6ARB333V (±0.1%, 10ppm/˚C)	33kΩ	33Ω	0.10%	8.3Ω	0.025%
InAmp Gain Error (0.015% ±35ppm/˚C)			0.02%		0.088%
InAmp Gain (Line 7 × Line 8)	**2.5**	**0.0029**	**0.12%**	**0.0028**	**0.113%**
V_{OUT} DM (Line 6 × Line 9)	**0V**	**5.5mV**	**0.22%**	**3.8mV**	**0.150%**
V_{OUT} CM (20V, 100db, ±1.5db over 0-50˚C)	**0V**	**200µV**		**37.7µV**	**0.038%**
V_{OUT} (Line 10 + Line 11)	**0V**	**5.7mV**	**0.23%**	**3.8mV**	**0.152%**

Table 4: 1A range

		At Nominal 25°C		Additional error over 0-50°C (Nominal ±25°C)	
Error	**Nominal Value**	**Abs. Error**	**Rel. Error**	**Abs. Error**	**Rel. Error**
Shunt Resistor: VMP-1R00-1.0-U (±0.1%, 20ppm/˚C)	1Ω		1%		0.05%
Input voltage error due to shunt	**1V**	**10mV**	**1%**	**0.5mV**	**0.05%**
Input voltage error due to switch leakage	**0V**	**4.5nV**	**0%**	**0nV**	**0%**
Input voltage error due to bias (I_{OS} ±0.5nA, ±20pA/˚C)	**0V**	**500nV**	**0%**	**500nV**	**0%**
InAmp: INA821 (V_{OS} ±35µv, 5µV/˚C)	0V	35µV		125µV	
InAmp Input Voltage error total (Sum of Lines 2-5)	**0V**	**10mV**	**1%**	**0.625mV**	**0.063%**
InAmp Gain Resistor R_G: ERA-6ARB333V (±0.1%, 10ppm/˚C)	33kΩ	33Ω	0.10%	8.3Ω	0.025%
InAmp Gain Error (0.015% ±35ppm/˚C)			0.02%		0.088%
InAmp Gain (Line 7 × Line 8)	**2.5**	**0.0029**	**0.12%**	**0.0028**	**0.113%**
V_{OUT} DM (Line 6 × Line 9)	**0V**	**28mV**	**1.12%**	**4.4mV**	**0.175%**
V_{OUT} CM (20V, 100db, ±1.5db over 0-50˚C)	**0V**	**200µV**		**37.7µV**	
V_{OUT} (Line 10 + Line 11)	**0V**	**28.2mV**	**1.13%**	**4.4mV**	**0.177%**

without being measured. This is the equivalent of under-reading the load current by 10pA, so it will result in a voltage error of up to 100nV (10pA × 10kΩ) at the op amp input.

The 610pA leakage will similarly cause a voltage error, but this time the error will be seen across the series combination of the shunt resistance and the switch on-resistance. This error will be 6.1µV (610pA × [10kΩ + 6Ω]). The total voltage error introduced by the switches will therefore be ±6.2µV, which you can see in line 3 of the error budget table for the 100µA range.

This is a meaningful amount compared with the instrumentation amplifier's ±35µV input offset voltage.

Given the relatively high shunt resistance, we also have to account for the impact of the instrumentation amp's input bias currents. The difference between these currents (the input offset current) will cause an additional voltage error across the source resistance. The INA821's data shows the maximum input offset current is ±0.5nA at 25°C, with a tempco (estimated from the graphs) of ±20pA/°C.

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This will result in a voltage error of ±5.0uV at 25°C with an additional ±5.0µV over the 0°C to 50°C operating range. This error, shown on line 4 of the error budget, is also similar in magnitude to the input offset voltage.

Other ranges

As you might expect from the above calculations, the error voltages will be lower for the other ranges where the shunt resistances are lower.

I went through the same exercise for these ranges and came up with error voltages due to switch leakage of 4.6nV for the 1A range and 95nV for the 10mA range, plus input offset current errors of 500pV and 50nV, respectively.

These are included in the relevant error budget tables (Tables 2-4), but are frankly so small as to be irrelevant given the instrumentation amplifier's ±35µV offset voltage.

The rest of the error budget tables are calculated as we did the last time. The upshot is a worst-case untrimmed 25°C error of ±1.13% for the 1A range and ±0.23% for the 10mA and 100µA ranges.

The big difference is due to the 1% tolerance of the 1Ω shunt compared to the 0.1% tolerance of the other two.

Over the operating temperature range, the 1A range has an additional ±0.18% error, with an extra ±0.15% for the 10mA and 100µA ranges.

Recall that the circuit in the previous article had a worst-case untrimmed 25°C error of 0.65% with ±0.28% additional error over temperature.

This circuit is better (except on the 1A range, where the shunt tolerance range has doubled) because we have used tighter-tolerance resistors and have reduced the instrumentation amplifier gain by a factor of 10.

Testing

As usual, I built the circuit and carefully measured its performance. The results are shown in the tables opposite (Tables 5-7). Again, we achieved much better performance than the worst-case calculations would suggest. The measured untrimmed errors were ±0.5%, ±0.06% and ±0.18% for the 1A, 10mA and 100µA ranges, respectively.

To calculate the trimmed error results, I used a gain correction based on the line of best fit, but just used the measured zero-current output value as the offset, mimicking the dynamic offset correction process. The trimmed errors were ±0.036%, ±0.054% and ±0.031% for the three ranges – very similar to the values we achieved previously.

The errors over the operating temperature range are around ±0.11%, assuming the offset calibration eliminates the offset component of the input-side temperature drift error. It would be around ±0.18% otherwise.

We can probably say that, across all ranges, our circuit achieves better than ±0.06% error at 25°C and ±0.25% over the operating temperature range. This is on par with the performance we saw last time, and means we have more-or-less met the expectations we set in Table 1 for these ranges.

As a paper exercise, I calculated the error budget for a possible 1µA full-scale range, assuming a 1MΩ 0.1% ±10ppm shunt. The worst-case untrimmed error at 25°C is ±0.35%, and the total error over the temperature range would be within ±0.6%, which is pretty good. With trimming, we could probably assume a current resolution in the order of ±5nA. This is about as low as I would go with this circuit.

Once we get down to measuring such small currents, things become very challenging.

A next obvious step will be to look into the analog-to-digital conversion process, to complete our theoretical PSU current-sensing design.

However, in all of our work so far, we have entirely ignored one important source of uncertainty and error: noise. This is an interesting but complex topic that we need to know about before moving on. So we will cover it in next month's article, before moving onto the digital side of measurement systems (ie, DACs & ADCs). PE

Measured Data		Untrimmed Error		Trimmed Error	
I (mA)	V_{OUT} (mV)	Absolute (mV)	Relative	Absolute (mV)	Relative
0.0000	-3.480	-3.48	-0.14%	0.00	0.000%
9.0707	223.366	-3.13	-0.12%	0.38	-0.001%
20.1549	500.500	-2.76	-0.11%	0.78	-0.007%
29.1271	723.712	-3.58	-0.14%	-0.02	0.006%
38.4297	955.808	-3.77	-0.15%	-0.18	0.016%
49.9203	1243.030	-3.46	-0.13%	0.16	0.018%
58.7674	1464.160	-3.24	-0.13%	0.41	-0.021%
72.2879	1801.780	-3.23	-0.13%	0.47	0.031%
80.1932	1998.150	-4.25	-0.17%	-0.53	0.000%
86.6674	2160.290	-3.77	-0.15%	-0.03	0.000%
95.1638	2373.240	-2.97	-0.12%	0.79	0.000%

Table 5 – 100μA range (V_{CM} = 20V).

Measured Data		Untrimmed Error		Trimmed Error	
I (mA)	V_{OUT} (mV)	Absolute (mV)	Relative	Absolute (mV)	Relative
0.00000	-0.773	-0.77	-0.03%	0.00	0.000%
0.98420	244.922	-0.83	-0.03%	-0.06	-0.002%
1.98602	495.514	-0.39	-0.02%	0.38	0.015%
2.93840	733.371	-0.34	-0.01%	0.43	0.017%
4.18878	1045.728	-0.20	-0.01%	0.56	0.022%
4.98283	1244.140	-0.06	0.00%	0.70	0.027%
5.85370	1461.660	0.01	0.00%	0.76	0.030%
7.11774	1775.860	-1.42	-0.06%	-0.67	-0.026%
7.99387	1996.360	0.31	0.01%	1.06	0.041%
8.68506	2169.050	0.42	0.02%	1.16	0.045%
9.53879	2382.450	0.64	0.03%	1.39	0.054%
10.64341	2658.070	0.44	0.02%	1.18	0.046%

Table 6 – 10mA range (V_{CM} = 20V).

Measured Data		Untrimmed Error		Trimmed Error	
I (mA)	V_{OUT} (mV)	Absolute (mV)	Relative	Absolute (mV)	Relative
0.000	0.055	0.05	0.00%	0.00	0.000%
100.303	251.400	0.95	0.04%	-0.27	-0.011%
199.851	500.786	1.76	0.07%	-0.61	-0.024%
300.618	754.046	3.41	0.13%	-0.14	-0.005%
400.330	1003.724	4.11	0.16%	-0.59	-0.023%
500.944	1255.870	5.03	0.20%	-0.85	-0.033%
601.552	1508.490	6.43	0.25%	-0.61	-0.024%
701.079	1758.470	7.90	0.31%	-0.31	-0.012%
800.656	2008.760	9.55	0.37%	0.18	0.007%
901.122	2261.360	11.29	0.44%	0.75	0.029%
1000.709	2511.350	12.61	0.49%	0.92	0.036%

Table 7 – 1A range (V_{CM} = 20V).

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Wide-Range OhmMeter	04109221	£12.95
JULY 2023		
Multimeter Checker/Calibrator	04107221	£10.95
JUNE 2023		
Arduino Programmable Load	04105221	£8.95
Buck-Boost LED Driver	16103221	£8.95
MAY 2023		
Precision AM-FM DDS Signal Generator	CSE211002	£14.95
Improved SMD Test Tweezers programmed PIC	0410621PIC	£12.95
AO 2x-dual-to-quad through-hole (pack of five)	AO1-MAY23	£6.95
AO 2x-dual-to-quad SMD (pack of five)	AO2-MAY23	£6.95
AO 2x-single-to-dual through-hole (pack of five)	AO3-MAY23	£6.95
AO 2x-single-to-dual SMD (pack of five)	AO4-MAY23	£6.95

PCBs for most recent *PE/EPE* constructional projects are available. From the July 2013 issue onwards, PCBs with eight-digit codes have silk screen overlays and, where applicable, are double-sided, have plated-through holes, and solder mask. They are similar to photos in the project articles. Earlier PCBs are likely to be more basic and may not include silk screen overlay, be single-sided, lack plated-through holes and solder mask.

Always check price and availability in the latest issue or online. A large number of older boards are listed for ordering on our website.

In most cases we do not supply kits or components for our projects. For older projects it is important to check the availability of all components before purchasing PCBs.

Magazine back issues are available – see https://www.electronpublishing.com/product-category/pe-back-issues-print-and-digital/

Double-sided | plated-through holes | solder mask

PROJECT	CODE	PRICE
APRIL 2023		
500W Amplifier Module	see p.22, April 2023	
Clipping Indicator (per channel)	01112211	£7.95
CD Welder Power Supply (one needed)	29103221	£9.95
CD Welder Controller (one needed)	29103222	£9.95
CD Welder Energy Storage module (several needed)	29103223	£7.95
AO Universal Dual Op Amp Board	AO1-APR23	£9.95
AO Stereo RIAA precision passives kit for dual op amp	AO2-APR23	£8.95
MARCH 2023		
Pico BackPack	07101221	£9.95
Semaphore Signal (controller)	09103221	£7.95
Semaphore Signal (blade)	09103222	£5.95
FEBRUARY 2023		
Tesla Coil driver board	26102221	£9.95
Tesla Coil potentiometer board	26102222	£5.95
Cooling Fan Controller & Loudspeaker Protector	01102221	£8.95
Remote Gate Controller	11009121	£12.95
JANUARY 2023		
Classic LED Metronome – 8-LED	23111211	£7.95
Classic LED Metronome – 10-LED	23111212	£8.95
Multi-Channel Speaker Protector – 6 channel	01101221	£9.95
Multi-Channel Speaker Protector – 4 channel	01101222	£7.95
Remote Control Range Extender – IR-to-UHF	15109212	£5.95
Remote Control Range Extender – UHF-to-IR	15109211	£7.95
AO Universal Single Op Amp Board	AO1-JAN23	£7.95
DECEMBER 2022		
Hummingbird Amplifier	01111211	£9.95
SMD Trainer PCB	29106211A	£8.95
SMD Trainer PCB + parts	29106211B	£13.95
NOVEMBER 2022		
USB Cable tester – main PCB	04108211	£12.95
USB Cable tester – front panel	04108212	£5.95
USB Cable tester – optional panel	SC5970	£5.95
Model Railway Carriage Lights – PCB	09109211	£6.95
AO trnsfmr PCB – standard VTX-A range	VTX-101-007	£6.95
AO trnsfmr PCB – dual-outline VTX102-3001/101-3002	VTX-Dual	£6.95
OCTOBER 2022		
SMD Test Tweezers – PCB and pair of tweezer arms	04106211-2	£11.95
SMD Test Tweezers – programmed PIC12F1572-I/SN	0410621A	£7.95
Tele-com	12110211	£12.95
SEPTEMBER 2022		
Touchscreen Digital Preamp – main board	01103191	£12.95
Touchscreen Digital Preamp – adaptor board pair	01103192	£5.95
20A DC Motor Speed Controller	11006211	£9.95
AUGUST 2022		
Multi-purpose Battery Manager – I/O Expander	11104212	£5.95
Multi-purpose Battery Manager – Switch Module	11104211	£8.95
Simple MIDI Music Keyboard (for 8 switches)	23101213	£6.95
Nano Pong	08105212	£7.95
JULY 2022		
Silicon Labs AM/FM/SW Radio	CSE210301C	£10.95
Level Crossing Controller	09108211	£6.95
JUNE 2022		
Full-wave Motor Speed Controller	1010221	£8.95
PIC Programming Helper for 8-pin PICs only	24106211	£7.95
PIC Programming Helper for 8, 14 or 20-pin PICs	24106212	£10.95
Advanced GPS Computer	05102211	£9.95

PROJECT	CODE	PRICE
MAY 2022		
Bus board PCB for Analogue Vocoder	AO1-MAY22	£10.95
Complete set of 14 PCBs for Analogue Vocoder	AO2-MAY22	£97.95
Programmed EEPROM for Digital FX Unit	FX1-MAY22	£10.95
Programmed PIC for Digital FX Unit w/ potentiometer	FX2-MAY22	£8.95
APRIL 2022		
64-key MIDI Matrix shield	23101211	£8.95
64-key MIDI Matrix switch board	23101212	£11.95
High-current Battery Balancer	14102211	£10.95
Digital FX Unit – using potentiometer	01102211	£9.95
Digital FX Unit – using BCD switch	01102212	£9.95
Universal Audio PSU	AO1-APR22	£11.95
MARCH 2022		
Mini Isolated Serial Link	24102211	£5.95
Busy Loo Indicator	16112201	£5.95
Analogue Vocoder – Band-pass filter board	AO1-MAR22	£9.95
Analogue Vocoder – HP/LP filter board	AO2-MAR22	£9.95
FEBRUARY 2022		
Arduino-based Power Supply	18106201	£9.95
Battery Monitor Logger	11106201	£10.95
Electronic Wind Chimes	23011201	£10.95
Analogue Vocoder – Driver Amplifier	AO-FEB22	£8.95

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All prices include VAT and UK p&p. Add £4 per *project* for post to Europe; £5 per *project* outside Europe.

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On-line Shop: www.electronpublishing.com

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NOTE: Most boards are in stock and sent within seven days of order receipt. Please allow up to 28 days delivery if we need to restock.

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PROJECT	CODE	PRICE
JANUARY 2022		
Vintage battery Radio Li-ion Power Supply	11111201	£9.95
MiniHeart: A Miniature Heartbeat Simulator	01109201	£8.95
DECEMBER 2021		
AM/FM/SW Digital Receiver	CSE200902A	£13.95
Balanced Input and Attenuator for USB CODEC	01106202	£11.95
NOVEMBER 2021		
Dual Battery Lifesaver	11111202	£6.95
OCTOBER 2021		
Mini Wi-Fi LCD BackPack	24106201	£8.95
SEPTEMBER 2021		
USB SuperCodec PCB	01106201	£14.95
Audio DDS Oscillator PCB	01110201	£5.95
Audio DDS Oscillator rotary encoder	01110201-ENC	6.95
Programming Adaptor Board for Audio DDS Oscillator	01110202	£5.95
High-power Ultrasonic Cleaner main PCB	04105201	£14.95
High-power Ultrasonic Cleaner front-panel PCB	04105202	
Night Keeper Lighthouse PCB	08110201	£5.95
AUGUST 2021		
Ol' Timer PCB	19104201	£11.95
Ol' Timer 8x8 RGB LED module using WS2812B	19104201-88	£8.95
Ol' Timer set of acrylic case pieces and spacer	19104201-ACR	£8.75
Ol' Timer DS3231 RTC IC wide SOIC-16	19104201-RTC	£5.95
Wideband Digital RF Power Meter	04106201	£9.75
Switchmode 78xx regulators ***(PACK of 5!)***	18105201	£7.95
Cool Beans SMAD display	CB-AUG21	£11.95
JULY 2021		
ATtiny816 Breakout / Dev Board w/ Capacitive Touch	24110181	£9.75
IR Remote Control Assistant (Jaycar version)	15005201	£8.95
IR Remote Control Assistant (Altronics version)	15005202	£8.95
PIC18F Development Board	PNM-JUL21	£12.95
Microphone Preamplifier	AO-JUL21	£11.95
JUNE 2021		
Roadies' Test Signal Generator (surface-mount ver.)	01005201	£8.95
Roadies' Test Signal Generator (through-hole version)	01005202	£9.95
Touchscreen Wide-range RCL Box (Resistor module)	04104201	£18.95
Touchscreen Wide-range RCL Box (Ind/Cap module)	04104202	
KickStart Part 3 – Gyrator-based Audio Filter	KS3-2021	£7.95
MAY 2021		
7-Band Equaliser (Mono)	01104201	£8.95
7-Band Equaliser (Stereo)	01104202	£10.95
Car Altimeter	05105201	£7.95
APRIL 2021		
Reflow Oven – DSP Active Crossover (CPU)	01106193	£19.95
Reflow Oven – DSP Active Crossover (Front panel)	01106195	
Reflow Oven – DSP Active Crossover (LCD)	01106196	
Frequency Reference Signal Distributor	CSE200103	£8.95
MARCH 2021		
Nutube Guitar Effects Pedal	01102201	£12.95
Programmable Thermal Regulator (Peltier Interface)	21109181	£18.95
Programmable Thermal Regulator (Peltier Driver)	21109182	
Tunable HF Preamp	CSE190502	£8.95
FEBRUARY 2021		
4G Remote Monitoring	27111191	£9.95

PROJECT	CODE	PRICE
JANUARY 2021		
Nutube Valve Preamplifier	01112191	£12.95
Arduino DCC Controller	09207181	£10.95
DECEMBER 2020		
Pseudo-Random Sequence Generator	16106191	£7.95
Clever Charger	14107191	£11.95
PE Theremin Amplifier	AO-1220-01	£8.95
NOVEMBER 2020		
LED Christmas Tree (1 off)	16107181-1	£6.95
LED Christmas Tree (4 off)	16107181-2	£14.95
LED Christmas Tree (12 off)	16107181-3	£24.95
LED Christmas Tree (20 off)	16107181-4	£34.95
USB/SPI Interface Board	16107182	£5.95
45V/8A Power Supply PCB *plus* acrylic spacer	18111181	£14.95
45V/8A Power Supply five-way display bezel	18111181-BZ	£3.95
Five-way LCD Panel Meter/Display	18111182	£7.95
OCTOBER 2020		
Digital Audio Millivoltmeter	04108191	£9.95
Precision Signal Amplifier	04107191	£7.95
SEPTEMBER 2020		
PE Theremin PSU	AO-0920-01	£5.95
PE Theremin PSU transformer	AO-0920-02	£7.95
Micromite Explore-28	07108191	£6.95
Ultrabrite LED Driver	16109191	£6.95
AUGUST 2020		
Micromite LCD BackPack V3	07106191	£9.95
Steering Wheel Audio Button to Infrared Adaptor	05105191	£7.95
JULY 2020		
AM/FM/CW Scanning HF/VHF RF Signal Generator	04106191	£13.95
Speech Synthesiser with the Raspberry Pi Zero	01106191	£5.95
PE Mini-organ PCB	AO-0720-01	£14.95
PE Mini-organ selected parts	AO-0720-02	£8.95
High-current 12V Battery Isolator – control	05106191	£6.95
High-current 12V Battery Isolator FET (2oz)	05106192	£9.95
JUNE 2020		
Arduino breakout board – 3.5-inch LCD Display	24111181	£6.95
Six-input Audio Selector main board	01110191	£10.95
Six-input Audio Selector switch panel board	01110192	
MAY 2020		
Ultra-low-distortion Preamplifier Input Selector	01111112	£11.25
Ultra-low-distortion Preamplifier pushbutton board	01111113	
Universal Regulator	18103111	£7.95
433MHz Wireless Data Repeater	15004191	£8.50
Bridge-mode Adaptor for Amplifier	01105191	£7.95
iCEstick VGA Terminal	02103191	£5.95
Analogue noise with tilt control	AO-0520-01	£7.95
Audio Spectrum Analyser	PM-0520-01	£8.95
APRIL 2020		
Flip-dot Display black coil board	19111181	£14.95
Flip-dot Display black pixels	19111182	
Flip-dot Display black frame	19111183	
Flip-dot Display green driver board	19111184	
MARCH 2020		
Diode Curve Plotter	04112181	£10.95
Steam Train Whistle / Diesel Horn Sound Generator	09106181	£8.50
Universal Passive Crossover (one off)	UPC0320	£12.50

Silicon Chip kits & parts for projects in this issue (posted from Australia):

ESR Test Tweezers complete kit	https://siliconchip.au/Shop/20/6952	AUD $60 (about £30) + P&P
Short-form kit for the Calibrated Microphone (SMD version)	https://siliconchip.au/Shop/20/6755	AUD $27 (about £13.50) + P&P
Short-form kit for the Calibrated Microphone (TH version)	https://siliconchip.au/Shop/20/6756	AUD $30 (about £15) + P&P

20% VAT is included in the above prices (it will be removed for customers outside the UK). Postage & packing cost is typically around £12-15. Please allow 1-2 weeks for delivery (longer in some cases).

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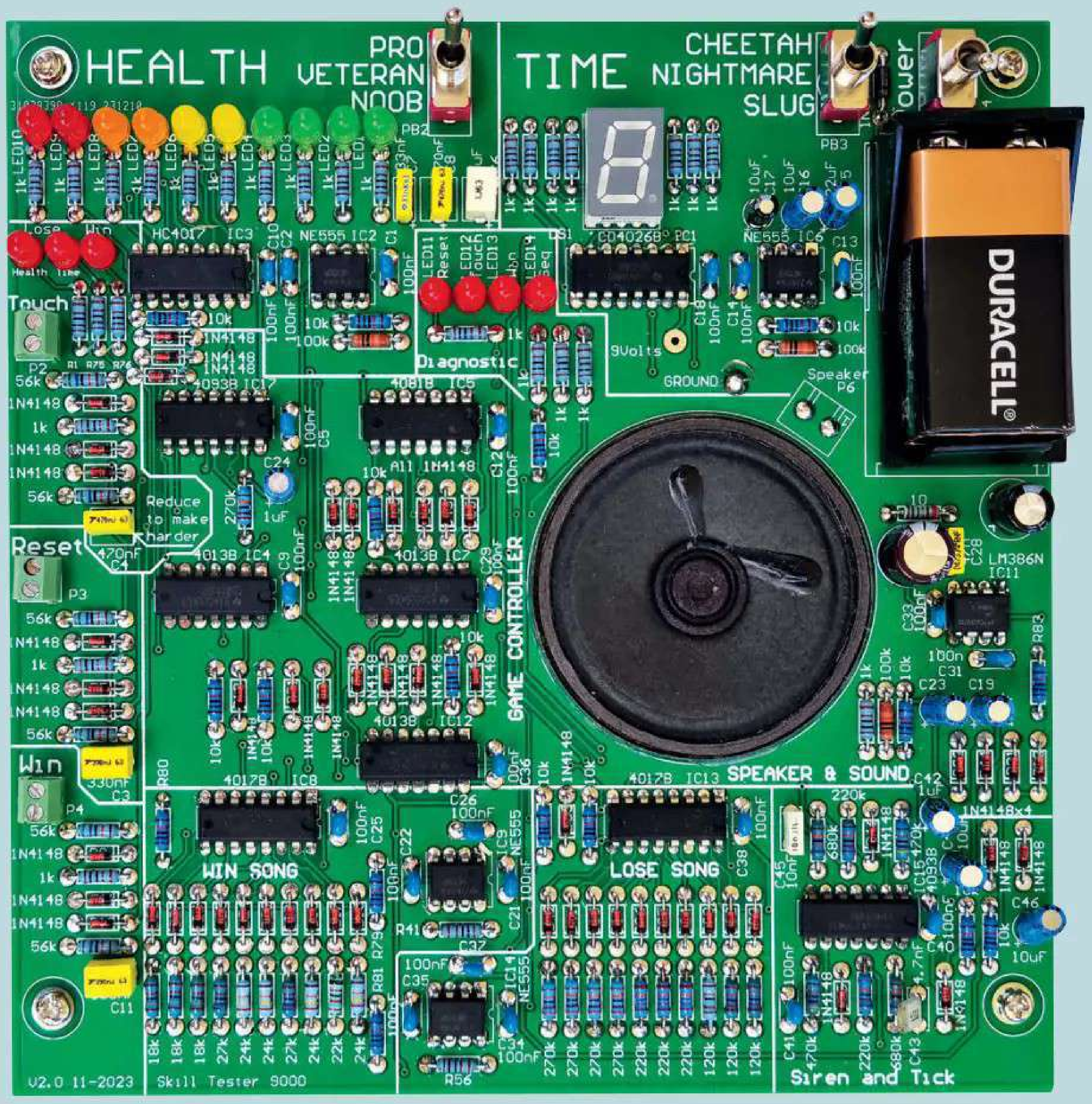
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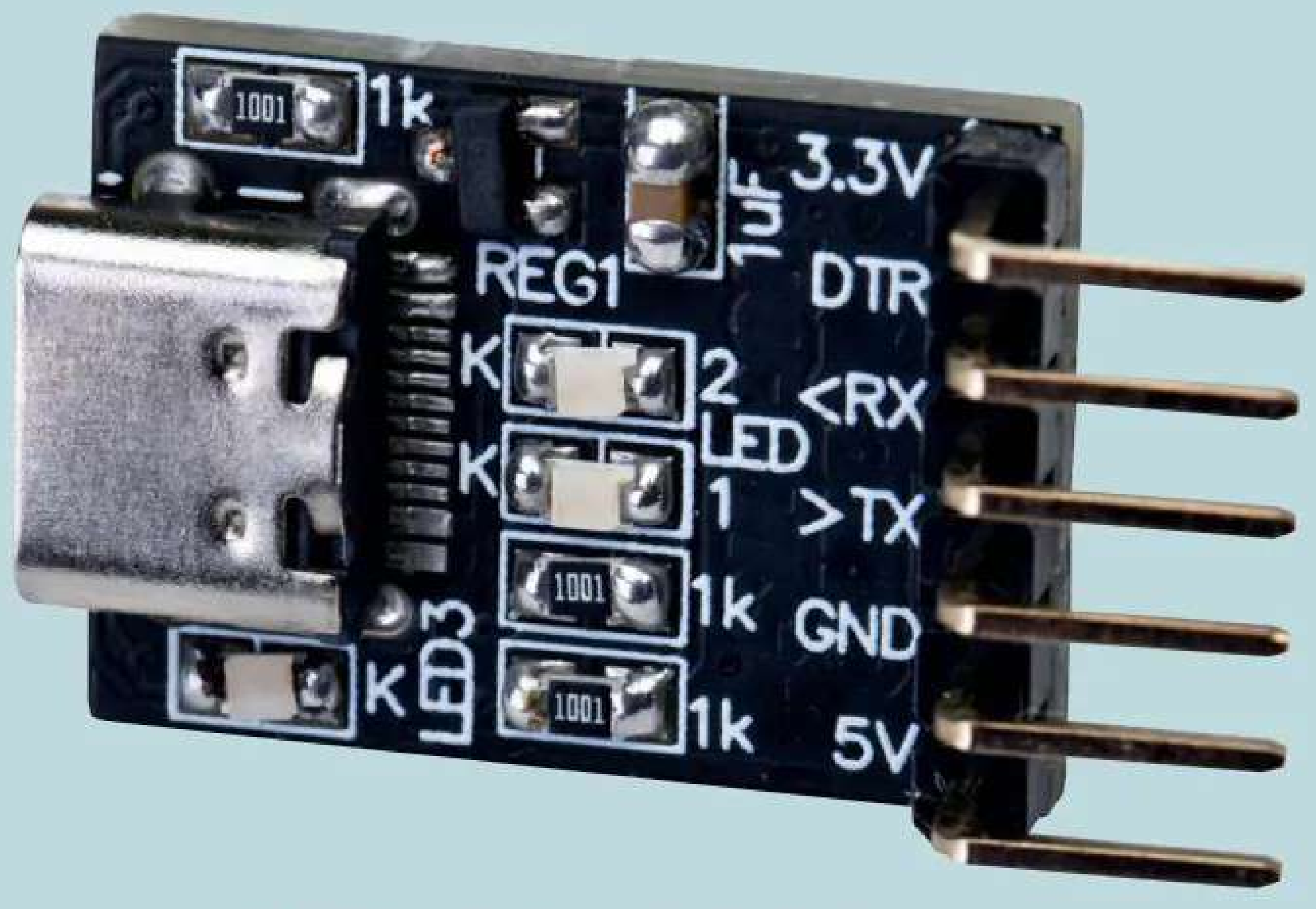
Siren and Tick
V2.0 11-2023
Skill Tester 9000

Skill Tester 9000

A fun variation on the wire and buzzer game – implemented with discrete logic!

(We ran out of space in this issue)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USB-C Serial Adaptor

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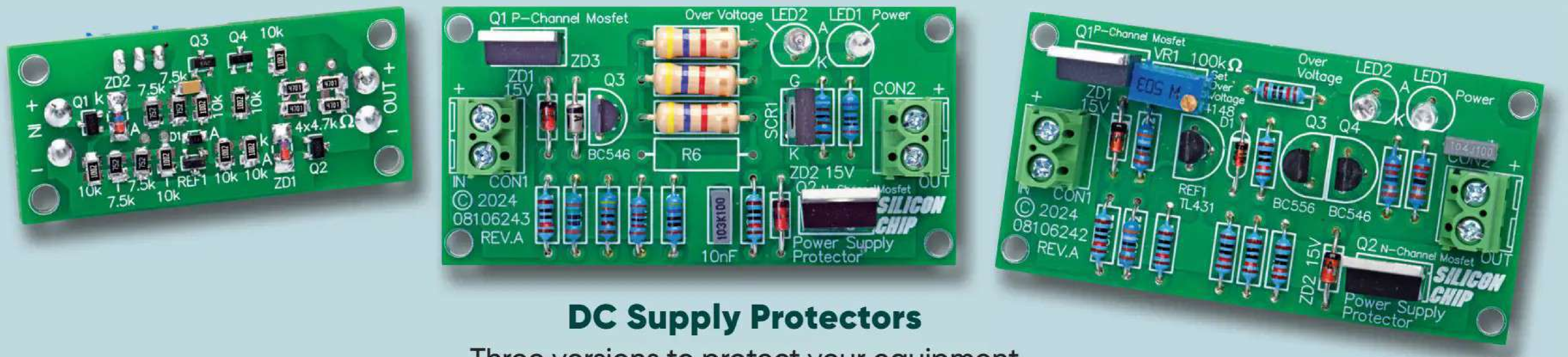
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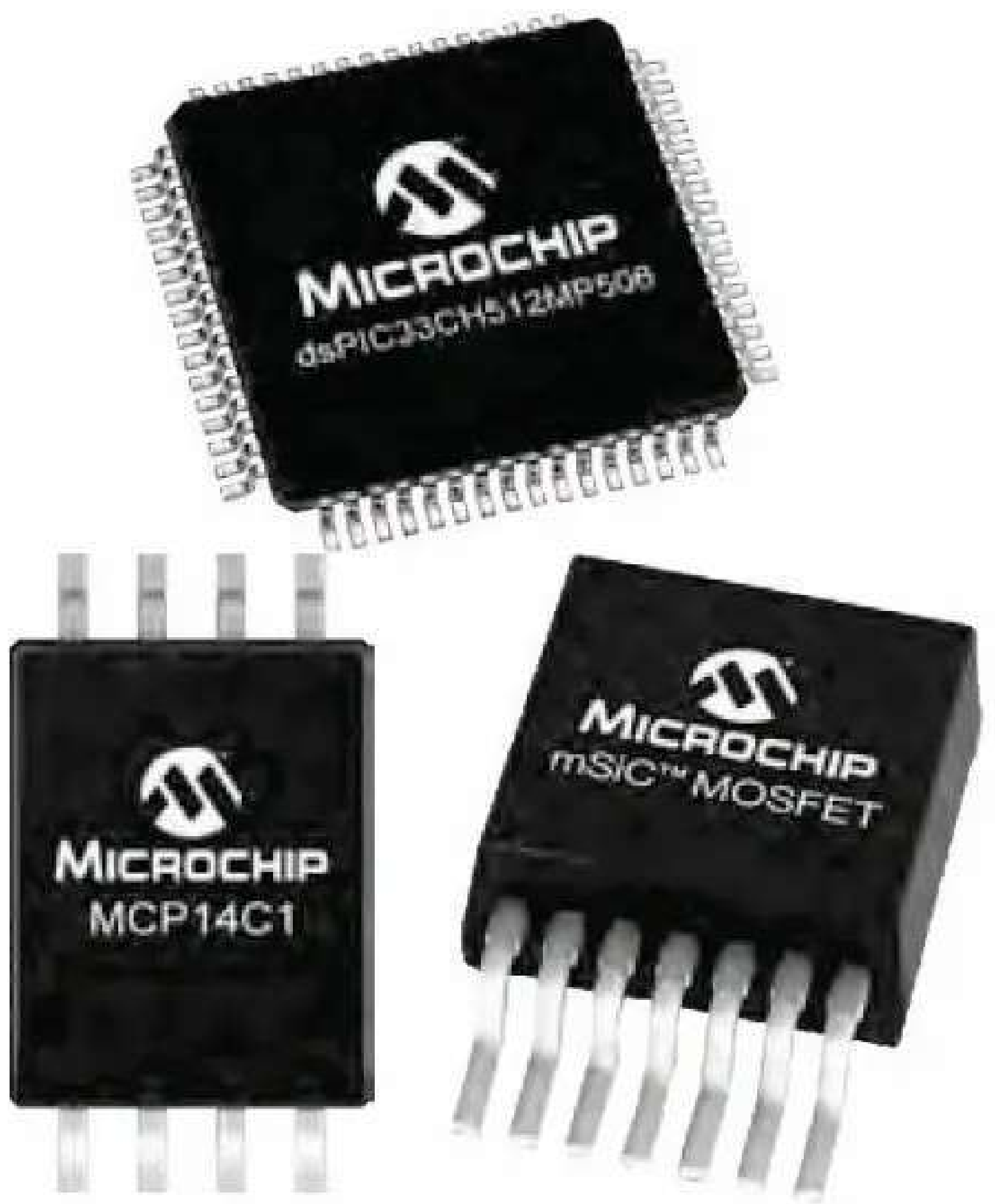

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